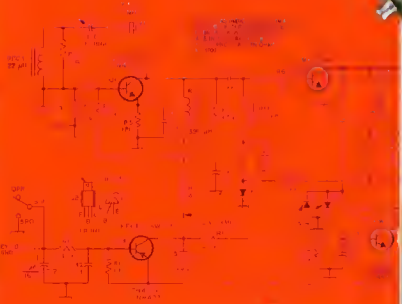
QRP CLASSICS

$12.00

The Best QRP Projects from QST and the ARRL Handbook



QRP CLASSICS

The Best QRP Projects from QST and the ARRL Handbook

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Edited by Bob Schetgen, KU7G

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Foreword

Welcome to the realm of QRP, a place where less RF power is more fun for the operator! In general, QRP operators use equipment that weighs less, takes up less space, costs less and is less dependent on ac power than the typical ham station of 50 W or more. In return, QRP enthusiasts get freedom—freedom to carry a complete station with accessories and antenna, in a briefcase. A typical QRP station is small enough to take along on vacation in a car full of family, by air or backpacking. Your QRP station can operate from batteries for long periods when the power fails, or indefinitely from unusual power sources such as private hydroelectric, wind or solar power systems.

Freedom is gratifying, but better still is the sense of accomplishment that comes from operating equipment you built yourself. You may best love the feeling when that first CQ from your home-built transmitter is answered—or the way a smile steals onto your face when that 1-kW station gives your 1-W transmitter a 599 report.

These qualities place QRP operation near the heart of Amateur Radio, and as a result QRP operation has always been a popular topic in League publications. In this book, we have assembled a balanced collection of QRP articles from 15 years of League publications for easy reference. While there are a few very simple projects suitable for beginners, you will also find challenging projects and pertinent articles about circuit design, component selection and adaptation. A QRP buffet is before you—enjoy!

David Sumner, K1ZZ
Executive Vice President

Newington, Connecticut
May 1990

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Preface

The QRP area of Amateur Radio is rich in experimentation, and the projects in this book cover 15 years of Amateur Radio technology. Feedback and pertinent work from the "Hints and Kinks" or "Technical Correspondence" columns of *QST* are included in this book. Parts availability changes every few months, however, and some parts mentioned may be difficult to acquire. In addition, many circuits use active devices well beyond their design range, and even the simplest circuits may not work with all samples of the listed components. So if a circuit doesn't work on the first try, don't let it discourage you. One joy of QRP is the relatively small investment of funds and time in most projects. You can try many circuit configurations and parts substitutions without breaking your budget.

Don't be frightened at the thought of experimenting with circuits; there are many resources to help. Some of the articles in this collection contain ideas that can be applied to other projects. Doug DeMaw, W1FB's *QRP Notebook* provides a basic foundation in QRP techniques with lists of standard component values and QRP organizations. Look in the *ARRL Handbook* for information about basic radio theory and circuit operation. Ask around, and you may find local hams with QRP construction experience. Your ARRL Section Manager (listed on page 8 of *QST*) can put you in touch with an Assistant Technical Coordinator near you who can help with technical questions and activities.

If you have trouble locating parts for projects, "Stalking Those Fugitive Components," in Chapter 2, should be of some help. Look for local parts suppliers in the yellow pages under "Electronic Equipment & Supplies - Dealers." Start building a library of manufacturers' data books to help you determine equivalent parts.

Finally, the QRP purist knows that the mode is officially limited to operation with less than 5 W (10 W PEP for SSB) of output power. ARRL extends the definition to include transmitters with up to 10 W of dc input power, regardless of transmitter efficiency. Some of the articles in this book significantly exceed either of these limits. Those construction projects that are not strictly QRP are included for educational purposes and for those cases where the operator decides to exceed the strict QRP limitations.

CALL FOR PAPERS

If there is sufficient support among QRP enthusiasts, we at ARRL Headquarters would like to publish a QRP Compendium. For such a book we will need a steady supply of previously unpublished articles. If you wish to contribute or want more information, contact the ARRL Technical Department for an author's guide. Articles should be clearly addressed to the Technical Department for the QRP Compendium.

From February 1990 QST, p 43:

Why QRP?

Low-power operation is more popular than ever before. Why not join in the fun?

By Kenny A. Chaffin, WB8E
2942 South Wabash Circle
Denver, CO 80231

Why would anyone except a masochist want to operate with less than 5 W output? What possible attraction could there be? Perhaps it's for the same reason *anyone* would operate an amateur station in this age of global telephone systems and satellite TV.

Maybe it's for the challenge of doing something a little different. Maybe it's for the thrill. But I can tell you, there's nothing quite like having a QSO with a Japanese, Russian, or rare DX station while running less power than a kid's nightlight!

The QRP Q signal was created to mean "Shall I reduce power?" but has since been adopted by the enthusiasts of low-power operation as their banner. QRP has come to mean 5 W or less output for CW, or 10 W PEP output or less for SSB. Most amateur organizations and contests embrace these as the official QRP limits.

Many of the same amateur activities that take place in the rest of Amateur Radio's domain are alive and well within the QRP community. These activities include constructing home-brew equipment, operating QRP stations, experimenting, DX chasing, and contesting.

You Can Build It

The QRP arena is one of the few places where the average home-brewer still can make a decent showing. In this age of multistage, integrated circuit, super-sophisticated all-mode transceivers, QRP operation stands out as a home-brewer's dream. How many hams can hope to duplicate the operation of the latest HF transceiver on their workbench? Probably none. If, however, we change the rules by restricting the power output, it is certainly possible for nearly anyone with the ability to obtain a ham license to build a 5-W transmitter.

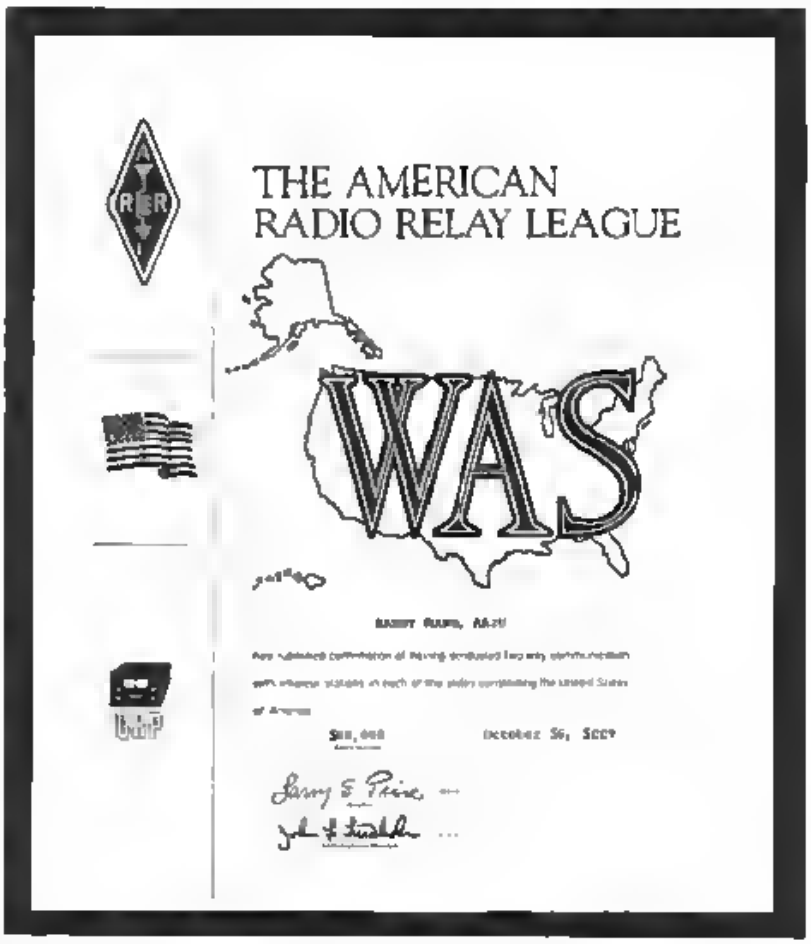
QRP transmitting equipment is simple and physically small. The same can't always be said for the receiver, however. A QRP receiver must do the same job as any other receiver, while usually in a smaller box. It is certainly possible to build an adequate QRP receiver by using minimal circuitry and integrated circuits—but it's not easy to duplicate a top-of-the-line commercial receiver in a matchbox.

If you are interested in home-brewing, but haven't actually done much, I would suggest the QRP transmitters as a good first project. QRP transmitters usually consist of a few transistors, and for HF work, the layout is not particularly critical. Probably the toughest part is finding or building the coils and chokes. Even the coils are not a big deal once you've wound a few. Schematics and kits are readily available. They make it easy to get started. After you've put together a kit or two, it'll be a piece of cake to move on to "bigger and better" projects.

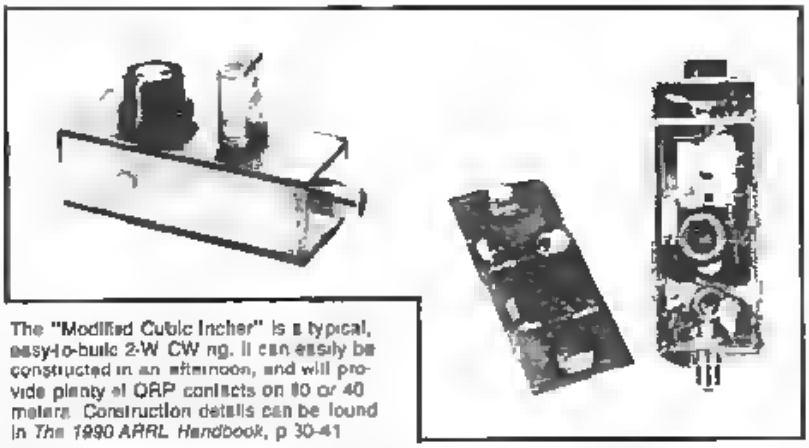
If you do start with a QRP transmitter, you can simplify the circuit even further by opting for crystal control. It may not be as restrictive as you think. A fair amount of QRP operation takes place on dedicated QRP frequencies—making it easy to pick the crystal you need (see Table 2). By adding a trimmer capacitor across the crystal you can "pull" the resonant frequency slightly to the lower side of the crystal frequency. (This is, in effect, a simple VXO circuit.) The crystal can be pulled from about 3 kHz on 80 meters to 10 kHz on 15 meters, depending on the crystal type and other factors.

Antennas

Once you have a working transmitter, you'll need a suitable antenna. Which brings us to the question: What kind of antennas do QRP stations use? You may think that following the lead of low-power, simple transmitter and receivers, QRP antennas should be small and simple. This is definitely not the case. A QRP antenna system should be as efficient as possible. Many transmission lines



ARRL's popular Worked All States award has a QRP endorsement. If you're looking for an interesting stateside challenge, why not break out a flea-powered transmitter and give it a try? To make things easier, we no longer require "QRP" to be indicated on the cards submitted for the award.



The "Modified Cubic Incher" is a typical, easy-to-build 2-W CW rig. It can easily be constructed in an afternoon, and will provide plenty of QRP contacts on 80 or 40 meters. Construction details can be found in *The 1990 ARRL Handbook*, p 30-41

attenuate the signal considerably before it reaches the antenna. If you have 5 W of RF output and a poor feed line, you could end up with only a couple of watts at the antenna! You should approach your QRP feed line as if it were being used for UHF or satellite work. You want to get as much power to the antenna as possible. Using a lossy feed line at kW power levels is tolerable; at QRP levels, however, the loss of every milliwatt becomes more critical.

The antenna itself is also important. For best results you need the best antenna you can put up—it's as simple as that—a high-gain Yagi if possible, up high and clear. It's just as though you were chasing the farthest DX. My antenna is a vertical, which is probably one of the worst choices. But it's the best I can do considering aesthetics, ordinances, and neighborly relations. Even with my vertical I've worked Japan and many Soviet stations using only 5 W output.

Books and Clubs

A couple of reference books you may want to pick up are, *The Joy of QRP* by Adrian Weiss, WØRSP, and *QRP Notebook* by Doug DeMaw, W1FB. The former is more operations oriented and the latter is almost entirely construction projects. There is also a monthly column on QRP written by Michael Bryce, WB8VGE, that runs in *73*. Another monthly QRP column appears in *World Radio*.

Occasional QRP articles, such as this one, appear in various Amateur Radio magazines. Several QRP clubs are available for those interested (see Table 1). QRP Amateur Radio Club International is one of the biggest, and publishes *QRP Quarterly*. For information about QRP ARCI and a sample copy of *QRP Quarterly*, write to Joe Sullivan, WA1WLU, 267 Sutton Street, North Andover, MA 01845. The Michigan QRP Club encourages low power operation with its newsletter, *The Five Watter*. And if you're interested in British-style QRPing, you can join the G-QRP club [G-land QRPing is strongly associated with home-brewing QRP gear.]

Operating Skills Required

If you want to hone your operating skills, QRP is for you. With only a few watts of signal to work with, it becomes mandatory to perfect your operating technique if you are

Confessions of an Inveterate Milliwatter

People like to overcome challenges; it's part of our nature. I lack the physical skills to be a mountain climber, so I have instead chosen to challenge the fickle layers of the ionosphere with a transmitter that runs milliwatts. It's my way of riding the knife-edge of what can be done.

Like many hams, I started chasing DX with 100 watts. I was content with this until a friend loaned me an HW-7. The meager 5 W didn't work very well with an indoor apartment antenna, but it gave me quite a thrill to work a few common European countries.

I finally managed to move to the country, where I had enough acreage to grow a better antenna crop. I also built a crystal-controlled transmitter that used a 74S00 logic chip as the oscillator and final amplifier, producing 250 milliwatts. A few local states were quickly put in the log. I smiled every time I told the station I was working that my final was a NAND gate!

A few hundred miles seemed to be the limit until the 1984 CW Sweepstakes weekend. I had never paid much attention to contests, so I was not prepared for the bedlam I found when I turned on my radio that Saturday afternoon. A loud W4 was calling CQ on 40 meters, and with no expectation of actually being heard, I sent my call sign—once. What's this? He's working me! Uh, let's see, I first got my ticket in, uh, '64—that will do. By the time the contest was over, I had worked 24 states with 250 milliwatts. Those big-gun contesters sure have good ears! Three years later, I had them all. My hand was literally shaking as I waiting for the band to improve enough to work a KL7 in the CQ WW contest.

Last year, my milliwatt quest continuing, I modified an HW-8 to run 10 milliwatts output. I had quite an adventure during the '88 CW Sweepstakes, netting 56 CSOs with 31 ARRL sections. The 18-hour operation boiled down to 347,200 points per watt!

The 1989 CW SS gave me state number twenty-nine. A couple of DX contests later, eight DXCC countries were in the log. All contacts were made via an 80-meter dipole fed with open-wire ladder line.

ARRL Lab Engineer Ed Hare, KA1CV, shows off his modified Heath HW-8 QRP transceiver. Ed's micro-power rig puts out slightly less than 10 milliwatts on 80 through 15 meters.

It's a high-tech effort. I use a computer to predict expected signal levels to those elusive western states. By all indications, WAS with 10 milliwatts can be done! If any operators west of the Mississippi want to test their station's weak-signal capabilities, I would appreciate a shot!—*Ed Hare, KA1CV, ARRL Lab Engineer.*

Table 1

QRP Clubs

QRP Amateur Radio Club International c/o Bill Harding, K4AHK 10923 Carters Oak Way Burke, VA 22015	Membership—$10 Newsletter—*QRP Quarterly*
Michigan QRP Club c/o Membership Chairperson 5346 W Frances Rd Clio, MI 48420	Membership—$7 Newsletter—*The Fine Watter*
G-QRP c/o George Dobbs, G3RJV 492 Manchester Road Roachdale Lance England OL11 3HE	Membership—$12 US Newsletter—*Sprat*

Table 2

Internationally Recognized QRP Frequencies (kHz)

CW	SSB	Novice
1810		
3560	3985	3710
7040 (7030)	7285	7110
10106		
14060	14285	
21060	21385	
24900	24950	
28060	28885	28110
50060	50885	

going to work through that DX pileup. QRP is the radio equivalent of brains over brawn.

But isn't a 1-W signal lost in the shuffle of more powerful stations? It's not as lost as you may think. A 1-W signal is only a little more than three S-units weaker than a 100-W signal. So, if your 100-W signal is S-9, your 1-W signal will be about S 6. And that's plenty of signal!

For QRP operation, you must be able to find DX stations, be aware of when and for how long bands will be open and have a crisp and clear setup on both CW and SSB. You must be able to quickly assimilate a DX operator's technique.

One of the primary skills QRP operation strengthens is patience. With QRP power levels you have to wait for the right moment and make your move. This means you must be alert and listening rather than transmitting. You have to be familiar with the bands, operating procedures of DX stations and other QRP operators. All this takes a bit of patience, practice and listening.

How Do I Do It?

Okay, let's say you just want to operate QRP without building any special equipment. That's easy, just turn the power down on your 100-W transceiver. This requires a power meter or some other method of determining your output power. This adjustment is dependent on your rig, and may be as simple as reducing the RF output control or as complicated as retuning the transmitter for reduced output.

Here's a neat experiment that will introduce you to the realm of QRP operation in a gradual fashion: cut your maximum output to half and operate at that power level for a week or so, then cut it in half again. Continue cutting power until you're down to 5 W. I'm sure you'll be surprised, as I was, at how well you can communicate with reduced power. In many cases, the operator on the other end can't tell the difference. My Heath HW-5400 puts out about 100 W maximum, and now that I work QRP almost exclusively, I really have to have a special reason to crank it up to full power.

Commercial QRP Equipment

If for some reason you can't operate your rig at reduced output, there is commercial QRP equipment available. Heathkit has offered three different QRP transceivers. All operate CW exclusively and cover only that portion of the HF bands. The first was the HW-7. It put out a few watts and had a relatively unstable receiver. The redesigned and improved version turned into the HW-8; there are plenty of these still in use.

The QRP community really took the HW-8 to heart and there are modifications galore available to spruce it up. Most of these have been collected in the *Hotwater Handbook*, available from Michael Bryce (he writes the QRP column for *73*). This handbook has been recently revised and reprinted, and includes mods for both the HW-8 and the latest-generation HW-9.

The culmination of Heath's QRP line is the HW-9. It features a vastly improved receiver and a bit healthier power output—slightly more than 5 W on some bands. The HW-9 also covers the newer WARC HF bands and is the only QRP rig currently on the market. You'll have to find the others at swap meets or through the classifieds. Expect to pay up to $70 for an HW-7, $60-$100 for an HW-8, and $100-$200 for a used HW-9.

The cream of the crop among QRP rigs is Ten-Tec's Argonaut series. The latest version (still long out of production) is the Argonaut 515. It's worth its weight in gold. The previously released 509 is almost as good and the 509's predecessor, the 505, is still hanging in there. These rigs operate both CW and SSB and are usually available at swap fests, through want ads, and from individuals. A 505 goes for $100-$175, a 509 for $125-$200 and a 515 for $200-$300 or more, depending on the market. Most of these rigs are generally available, it's just a matter of whether you can afford, and find, a 515 or an HW-8.

A Few More Advantages

There are a couple of other advantages of QRP operation that aren't so obvious. Because you are operating with a minimal power output, your transmitter will probably last "forever." Your electric bill will be less—especially if you stop using your 2-kW space heater. The other nonobvious advantage is that you won't overload the front end of your neighbor's television. It's a pretty rare occasion when operating with 5 W causes interference.

Contests and Awards

The bonus multipliers and points for QRP contest operation have gotten many hams hooked on QRP. Operating "QRP battery power" for Field Day gives a multiplier of five. You only have to make one contact for every five QRO QSOs.

QRP operation is becoming quite popular for many major contests. The following contests have QRP categories: November Sweepstakes, June and September VHF QSO Parties, January VHF Sweepstakes, and the ARRL International DX Contest, among others.

As far as awards, QRP ARCI offers a thousand-miles-per-watt award, available to anyone presenting evidence of a qualifying QSO. QRP ARCI also offers special QRP awards for WAS, WAC and DXCC. The other QRP clubs also offer versions of these, and other, QRP operating achievement awards.

What's Left

What do you do once you've completed QRP DXCC? How about milliwatting? Milliwatting is operating at less than 1-W output. Once you've perfected your QRP skills and equipment, this is the next challenge. Admittedly, there are few who strive for these ranks, but when it all works—WOW! I've recently seen a circuit for a half-watt crystal-controlled transmitter using a single 2N2222 transistor. I haven't tried it yet, but when I do, I can't wait to hear what the operator on the other end says when I tell him. Of course, at milliwatt levels your antenna and feed line become doubly critical. It seems strange to see a 1-inch-square, single-transistor transmitter connected to ¾-inch hardline. But it's great fun.

So why not give QRP or milliwatt operation a try? You just ought get hooked. See you on 7040 kHz—a popular QRP hangout.

From April 1984 QST, p 52:

QRP: More Than a State of Mind

Looking for a new challenge? Try reducing power and adopting a few new operating habits.

By Bradley Wells,* KR7L

Low-power operation, or QRP, has enjoyed a surge in popularity in recent years. Why? Mostly it's the challenge of working stations the "hard way," be it during contests or everyday operation, and the great satisfaction that comes from making contacts that the "big guns" make. Most low-power ops will agree that the motivation for QRP is the same as for chasing DX — but the rewards are inversely proportional to the amount of power used.

In this article, we'll take a look at the exciting world of QRP, discuss some equipment that's available and talk about ways of improving your chances of success with low-power operation. One word of caution to the reader though: QRP can be habit-forming.

The definition of QRP, recognized by most amateur organizations, is 10-W input, or 5-W measured output. Five watts may not sound like much to those who consider 200 W low power, but the difference is not as great as you may think. Under actual conditions, 5 W will have little effect on your ability to work DX. The difference between QRP and, say, 200 or 2000 W is only 3 or 5 S units. Also, QRP exemplifies the spirit of the Rules — specifically 97.67(b), which states that " . . . amateur stations shall use the minimum amount of transmitter power necessary to carry out the desired communications."

Choosing an Antenna

A major failing of both experienced and novice QRPers is the antenna system. Unfortunately, most hams think low power equates with poor antennas. Many QRP operators seem to delight in using their rig with a 50-foot piece of wire thrown out the nearest window.

The basic rule of QRP antennas is that nothing beats a beam; and nothing beats a beam on a tall tower. Put up the best beam/tower combination you can afford. A good 3-element beam and 40-foot tower will put you on a more-than-equal footing with those running 200 W to a vertical.

A good full-size dipole is the next best choice. On 20, 15 and 10 meters, a high dipole exhibits directivity, so place it broadside to the desired direction of radiation.

Related to the dipole, and almost as easy to construct, is the single-quad loop. This antenna is more directive, has wide bandwidth and can exhibit up to 2-dB gain over a dipole.

The poorest choice for the QRPer is the vertical antenna. The vertical suffers two defects when compared to a dipole. It is highly susceptible to man-made QRN, notably power-line noise. For a vertical to have the same radiation efficiency of a dipole, a good radial system is required. Amateurs lacking space for beams or dipoles might consider the Cushcraft R-3 tuned vertical, which requires no radials and approaches the efficiency of a half-wave dipole.

Do not skimp on the coax. Use the best grade of RG-8 you can afford. We are not interested in power capability, but in achieving the lowest attenuation possible. The ham with an amplifier will not miss a couple of watts heating his coax as much as the QRPer running 5 W will. For portable operation, RG-8X may be used where its light weight and ease of handling offset the increase in attenuation. Make all connections clean and weatherproof. Strive for the highest possible efficiency in both feed line and the antenna.

Operating Tips

One may wonder how a DX station can hear a 5-W signal when megawatts are coming at him. But hear it he does, and more often than not the experienced QRP operator will get through those pileups to snag the rare DX station. To do this, however, the operator requires some knowledge of tactics used by successful stations.

First, and most important, listen before using your key or mic. Is he working stations by call area or at random? Is he picking up callenders? Is he listening high or low, and how wide is the split? All of these things can only be learned by listening. Spend five, even 10 minutes on your receiver before you begin to transmit.

Second, invest in a memory keyer. You're going to send your call a number of times, and it's much easier to do so by pushing a button instead of wearing out your wrist. Send your call at a slightly slower speed than the DX station is transmitting.

Third, on phone, use standard phonetics. The ham on the other end doesn't have time to figure out cute call signs, and will ignore you. In addition, use some form of speech processing to boost your average power, but don't overdo it. Too much is far worse than too little.

Fourth, time your calls. This is most important for QRP operators. Don't try to be first to hit the keyer or PTT switch. Normally, everyone will send their calls all at once, pause, then try again. When you hear that pause, slip your call in just once. That's all you have time for. Do this correctly, and you may get through on the third or fourth call.

Finally, know when to quit. Everyone has days when the propagation is wrong or Lady Luck is against you. Believe it or not, the world will not end if you fail to work the DX in that pileup.

It's for Success

With only 5 W, there is no way you're going to blast an opening into a crowded band. You don't have an "afterburner" to klek in under heavy QRM conditions, or the power to make your own propagation. So, you need a change in operating style.

The first habit you will break, and soon forget, is calling "CQ." In fact, "CQ" and "CQ DX" will just about disappear from your vocabulary and keyer. With full legal power, a "CQ" in any direction will get you contacts. QRP will never bring the same results. For those unwilling to change this operating habit, the kiss of death is on their QRP career.

There are several ways to increase your chances of success. First, have a good beam antenna. Second, sign your call with /QRP. This may cause stations to call you out of curiosity. The idea is to let everyone know, up front, why you're not 40 dB over S 9. However, most hams will not answer a weak "CQ" unless your call begins with something like S79, VK0 or T32.

The single-most-effective QRP operating technique is search-and-pounce. Search-and-pounce is simply tuning carefully through each band until you find a station to work. Most of the stations you work will

*1290 Puget Dr E, Port Orchard, WA 98366

be calling "CQ," or you will nail them as they finish a QSO.

Work the station with a moderate-to-loud signal. Since the sensitivity of most QRP receivers outstrips the effective range of their transmitter, a signal that is very weak may be impossible to work. Propagation is a reciprocal thing, and if the station on the other end is S 1 running a kilowatt, imagine what 5 W will sound like. Actually, there will be no sound at all — you simply will not be heard. This condition is more prevalent on 80 and 40 meters, where antennas and propagation tend to work against the QRPer.

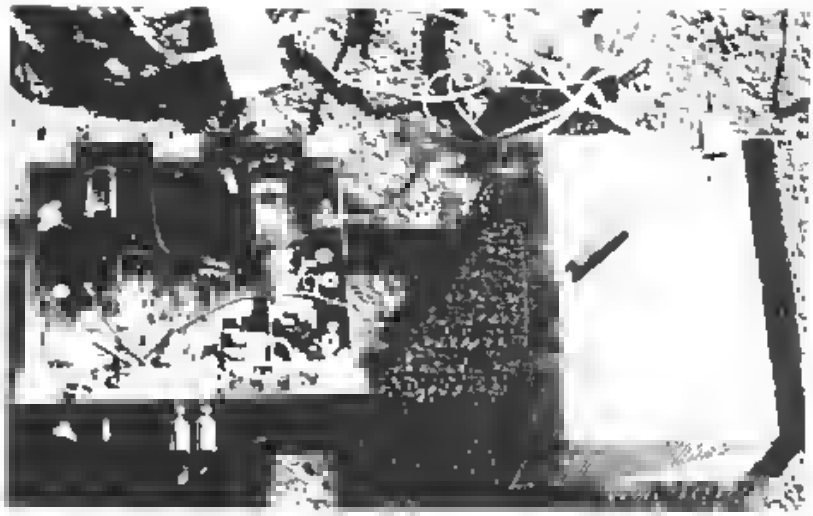
If you become involved in a marginal contact, don't prolong it. The other operator did you a favor by coming back and will not get much enjoyment out of the QSO if you're only 3 3 9 at his end. The place to tell him all about your rig, antenna and the weather is on your QSL card.

A fact of QRP life, and one of its more frustrating aspects, is that you are going to get stomped on occasionally — whether it's deliberate bad manners, carelessness or simply that the station firing up on frequency can't hear you. Sometimes, you can operate through the QRM, but generally it's the end of the QSO.

For those of you who chase DX (and who doesn't?), listening on the local DX repeater is a good way to expand your search-and-pounce technique. If you do spot a bit of DX, work him first, then announce his frequency over the repeater. Do it the other way around and you may find yourself hip-deep in "big gun" stations.

Another prime requirement for being able to work DX (or anyone else) on a consistent basis is at least a working knowledge of propagation. All of the major amateur publications have monthly propagation charts. They use different formats, so different interpretive techniques are applicable to each. All of these charts are prepared several months in advance of publication; you should be able to update their information to make allowance for current conditions. There are two ways to do this. One is to monitor the WWV propagation forecast at 18 minutes after each hour. These recordings provide real-time information to update your monthly charts. A second method is to subscribe to one of the DX bulletins. Printed on a weekly or biweekly basis, all are excellent indicators of relatively current propagation conditions.

The three bands providing the bulk of activity for QRP are 20, 15 and 10 meters. When the 10-meter band is open, there is little difference between 5 and 500 W. It can exhibit rapid shifts in propagation, however, which can be disconcerting to even experienced hams. Twenty meters is the most consistent band, providing openings to some part of the world day and night.



Jim Ford, N6JF, of Costa Mesa, California, went the low-power route, and he's glad he did. Using the QPS "Hamcation Rig" (April 1983 *QST*) as a guide, Jim built his own QRP rig which he operates with great success on 10 MHz.

Forty and 80 meters are less consistent producers because of their more-seasonal nature and higher levels of QRN and QRM. Both tend to be winter bands, but can produce results any time of year. The best DX time is 30 minutes before and after local sunrise or sunset. Also, the 30-meter band is excellent for QRPers. Its propagation lies midway between 20 and 40 meters, and only limited-power (250 W) operation is permitted.

Most QRP CW operation is around 40-60 kHz up from the bottom edge of any band. Most phone operation tends to be in the Advanced and Extra Class subbands. Stay out of the Novice segments; beginners have enough problems without the added difficulty of having to copy less than S 9 signals.

The QRP Contester

For many, contesting is just one interesting facet of Amateur Radio. For others, contests *are* Amateur Radio. Non-contesters and contesters alike may view operating a contest with a QRP rig as the ultimate insanity. Actually, the reverse is true. Most of us don't have the megabucks required to put together a top-drawer, big-gun, killer-type contest station. However, most hams can afford a first-class QRP station. Since QRP rigs are relatively inexpensive, you can afford to invest more in antennas — a deciding factor in contesting.

Many contests have a separate single-operator, all-band QRP category. Thus, you need only compete against other QRP operators. However, winning still requires maximum doses of perseverance and a large amount of skill.

Contesting effectively with QRP requires the application of several important techniques. At the beginning of the contest, work the strongest stations. Then, work the progressively weaker stations. In addition, don't waste too much time calling any one station. If he hasn't come back to you by the fourth call, move on. You can work him later when the pileup is reduced. An exception to this would be near the end of the contest when that DX station represents a new multiplier.

Instead of tuning up and down the band, start at the high end and work stations as you go to the low end. When you hit the bottom edge, quickly tune up to the top and start down again. This will maximize your time on all portions of the band. Those proficient with a search-and-pounce technique will have a QSO rate almost equal to most stations calling "CQ." Also, new stations will appear and disappear with great rapidity, so don't worry about working the band dry.

Another rule for the QRPer is to work the MUF (maximum usable frequency). Work the highest frequency that is open in the area you want to cover, based on WWV or other propagation information. Operating at or close to the MUF reduces path loss and maximizes your 5-W signal.

In a DX contest, know the areas that are easiest to work, and concentrate on those at the start of the contest. Work the more difficult areas during the last 24 hours. For example: Generally, Japan, Oceania and Europe can be worked from the West Coast on 20 meters in the morning. For the QRPer, however, it is more productive to work Japan and Oceania Saturday morning and Europe Sunday morning. By the last day, Europeans will have worked out much of the Eastern seaboard and will respond more quickly to a call from the West Coast.

In any contest, but more particularly in a DX contest, establish some type of game plan. Spend some time consulting propaga-

tion charts, and write up a time-versus-frequency plan for your own use. Decide which areas you will cover at what times and the best band for each combination. This plan should be used as a guide for each hour of operation. The most productive directions will be based on your experience and an examination of previous contest scores.

Next to your log, the most important record to keep is the dupe sheet. Duplicating contacts means wasted effort, lost points and less-productive operating time. Since, as a QRP station, you will be operating 99% of the time in a search-and-pounce mode, your dupe sheet must be as current as your contest log. There are as many different dupe sheets as there are contests, so use one that fits your needs.

Finally, keep the proper perspective and attitude before, during and after the contest. Above all, don't worry about the big-gun station down the block. You're not competing against him, only against other QRPers.

From September 1981 QST, p 11:

Experimenting for the Beginner

Experimenting is half the fun of Amateur Radio! QRP (low power) gear is great for the newcomer to this fine art. Here's how to get started.

By Doug DeMaw, W1FB
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What's this? You've never built a piece of amateur equipment? You don't know anything about circuits, so you just operate? Well, if this description fits you, at least half the thrill of being a ham has eluded you! For many of us the greatest excitement in amateur work came from building and using that first transmitter. There's a special feeling connected with telling the other guy or gal, "The rig here is homemade." If you haven't been able to make the statement over the air, perhaps it's time you did!

Most experimenters start out with relatively simple projects, and rightfully so. In the old days some of our tinkerers enjoyed building one-tube transmitters. Often, the name of the game was "power output." That is, we tried to extract more output power from a single oscillator than the tube was designed to deliver. A number of popular transmitters of this type were described in *QST* by F. Sutter.[1] But today it's prudent to use transistors and to operate them within their safe maximum ratings. QRP equipment (generally 5 watts or less of rf output power) can provide many interesting and exciting hours of operation, and it's easy and inexpensive to build. Therefore, QRP is the theme of our article this month on basic radio learning.

How to Experiment

We need not have college degrees in engineering to conduct experiments in nonprofessional electronics work. We can assemble suggested circuits, test them, learn their characteristics, and then make changes and observe the results. Familiarity with fundamental circuits can lead to circuit improvements and innovations, and perhaps later to some original design work. Many of the early-day inventors of electrical and electronic devices and systems followed this approach, which supports the validity of the precept, "Learn by doing."

We amateurs have the advantage of trying our ideas at home rather than at work. So, if the circuit is a flop, no need to contemplate the unemployment line! Furthermore, if the equipment is a transmitter for one of the amateur bands, we are licensed to put it on the air and to give it a true "environmental test," an advantage not enjoyed by many engineers and technicians.

The simplest approach we can take to experimenting is to adopt the breadboarding technique.[1] This allows us to hack a test circuit together quickly and easily. In the process we cut down on expense and eliminate the chore of laying out and etching a circuit board. The final product may not look like a work of art, but it can be used on the air just as effectively as a commercial-looking version of the same circuit.

Bargain-bag assortments of 1/4- and 1/2-watt resistors are a vital part of the experimenter's workshop. Likewise with assortments of disc ceramic capacitors, trimmer capacitors, volume controls and small electrolytic capacitors. Of course, we need a small pencil type of soldering iron (40 watts), some solder and a few feet of light-gauge, insulated hookup wire. Bargain assortments are often available from Radio Shack, Poly Paks and other prominent vendors. The best deals are often available at Amateur Radio flea markets, so we must be on the alert when browsing at hamfests and conventions.

An important item in our workshop is a VOM (volt/ohm/milliampere meter). Even a low-cost imported instrument will suffice if cost is an important consideration. For rf measurements it is wise to have a VOM that can be used with a

[1]Footnotes appear at end of article

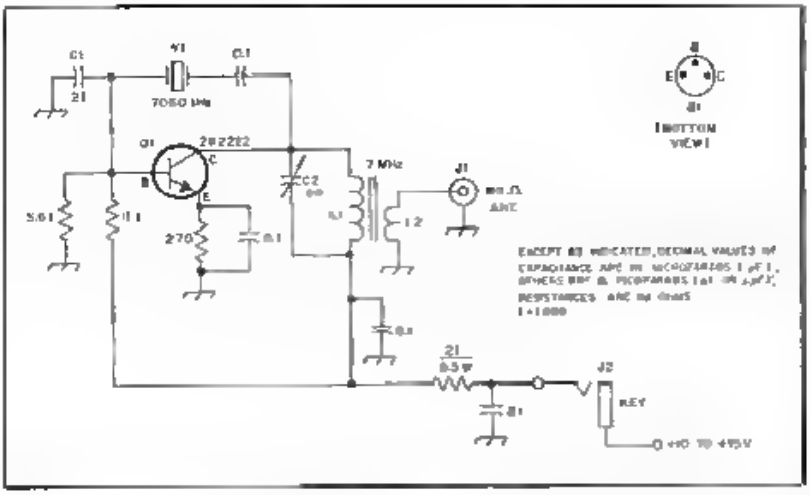


Fig 1 — Circuit of a one-transistor QRP transmitter. Fixed-value capacitors are disc ceramic, 50 volts or greater. Resistors are 1/4 or 1/2-watt composition, 10% tolerance. C1 described in text. C2 is a 100-pF mica trimmer. L1 is a 6-µH winding of 34 turns of no. 26 enam. wire on an Amidon or Palomar T50-2 toroid core. L2 is 8 turns of no. 26 enam. wire wound over L1 winding (see text). J1 is a phono jack, and J2 is a 2-circuit phone jack. Y1 is a fundamental surplus or new crystal for the standard 40-meter QRP frequency (7060 kHz).

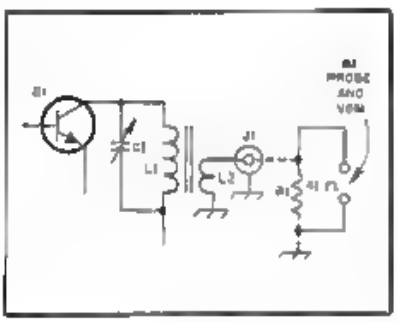


Fig 2 — Details for measuring transmitter output power with a dummy load (R1), an rf probe and a VOM (see text).

homemade rf probe.[1] This will permit us to measure rf voltages in oscillators and transmitters when performing initial checkout or debugging. A frequency counter is very useful to the experimenter, and should be acquired if the expense can be justified.

We will need a dc power supply for our workbench, and for most of our experiments we can manage nicely with a 12-volt, 1-ampere regulated supply.[2] If the output voltage can be made variable, so much the better.

Bargain assortments of transistors, ICs and diodes aren't likely to be of much use to us unless we have a way to locate the defective ones. Most "bargains" of this type contain manufacturer's rejects, and 50% or more of the semiconductors in a bag are often open, shorted or leaky. Therefore, we're better off to buy parts of known quality for each of our experiments. This practice will help us to avoid confusion and despair.

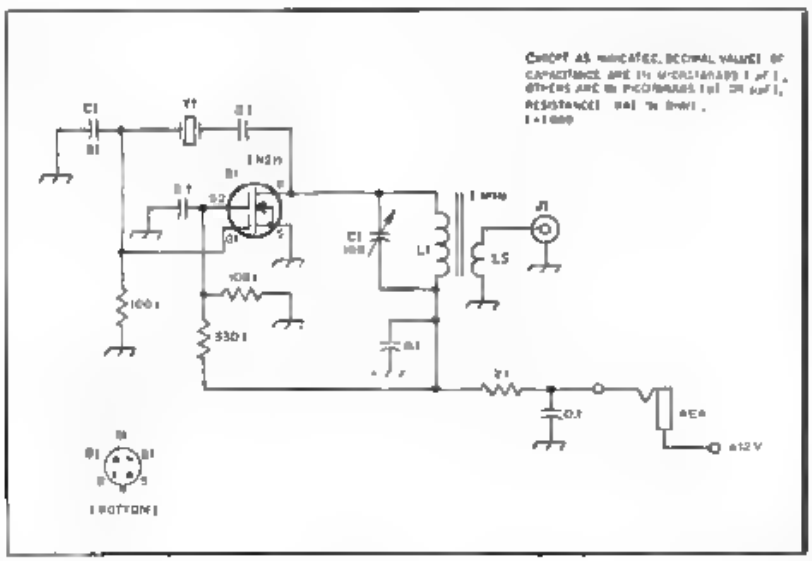


Fig 3 — Same circuit as Fig 1 except that an FET is used at Q1.

The Simplest Transmitter

How uncomplicated can a transmitter be for experimental work? Factually, a one-transistor oscillator qualifies as a transmitter. Many beginners have had exciting results with such a circuit while operating with only 50 milliwatts (0.05 watt!) of power output. For example, the circuit in Fig. 1 was tacked together one lunch hour in the ARRL lab and was connected to a 28-foot (8.5-m) base-loaded vertical antenna with buried radials. On the third CQ an answer came from a W8 in Ohio. A signal report of RST 569 was received for our 50-mW signal on 7060 kHz. A second QSO with a W2 station in New Jersey netted an RST 589 report!

Y1 of Fig. 1 determines the operating frequency. C2 tunes L1 to the approximate frequency of Y1. If it is set for resonance at exactly 7060 kHz in this example, the cw signal may become chirpy. With this type of oscillator it is best to tune the C2/L2 circuit for the best sounding note consistent with reasonable power output. Maximum power will not coincide with the cleanest cw note when connecting an antenna to this type of oscillator unless very light coupling is used (L2) between the tuned circuit and the antenna. The lighter coupling will, in itself, reduce the available power to the antenna.

The circuit of Fig. 1 can be used on 160, 80, 40 or 20 meters by using a fundamental-cut crystal for the desired frequency. C1 is part of the feedback network and will have to be chosen for the crystal we use. This is because some crystals are more active than others. The more sluggish a crystal is, the greater the feedback voltage required to make the circuit oscillate reliably. Values between 15 and 100 pF are typical for use at C1 in this particular circuit. We can experiment with the number of turns in L2 to extract maximum rf power output from the circuit.

Fig. 2 shows how we can use a 47-ohm resistor as a dummy load to measure the output power. An rf probe (mentioned earlier) and VOM are connected across R1 with the key closed. Output power can be calculated from,

$$P = E^2/R$$

where P is in watts, E is in rms volts and R is in ohms. Therefore, if we measured 1.53 volts across R1, we would have an output power of 50 milliwatts (0.05 W). The accuracy of our measurement depends on the purity of the sine wave from the transmitter. A distorted waveform will

yield only approximate power-output readings on the VOM. A 51-ohm resistor could be used at R1, but that is a 5% tolerance (gold band) value, and would cost more than a silver-band (10% tolerance) resistor. So, we can use a 47- or 56-ohm resistor. Either value is close enough to 50 ohms for our purposes. Here again is an example of the joy of *experimenting* versus designing!

We can also use field-effect transistors as oscillators of the kind illustrated in Fig. 1. The version seen in Fig. 3 contains a dual-gate MOSFET. Output power from this circuit will be somewhat lower than that from the bipolar transistor oscillator of Fig. 1, but plenty of QSOs can be had with this simple transmitter. Other dual-gate MOSFETs could be used in place of the 3N211, such as a 40673.

If we decided to use a VFO to control the operating frequency of the transmitter in Fig. 1, we could make the modifications shown in Fig. 4. Y1 and C1 are removed to prevent oscillation at the crystal frequency. A dc-blocking capacitor (C3) is added as shown. The rf voltage (rms) developed from the base of Q1 to ground (with the VFO connected and operating) should be between 1 and 3 volts for best results. This shows just another way we can experiment with simple circuits.

Additional experiments can be conducted with the one-transistor transmitters by trying various types of transistors in the basic circuits of Figs. 1 and 3. One important transistor characteristic is the maximum operating voltage (V_{ce}), which should never be rated less than two times the supply voltage for your work. This will allow for the voltage swing (peak to peak) during the rf sine-wave cycle at the collector or drain. If the voltage is allowed to rise beyond the specified safe value, the transistor can "go away" instantly! We must be concerned also with the upper frequency rating of the semiconductor. This is usually specified as f_T. A good rule of thumb for obtaining maximum oscillator or amplifier performance is to use a transistor that has an f_T at least five times higher than the chosen operating frequency. Thus, for 7-MHz operation the f_T should be 35 MHz or higher. Most FETs are rated for a maximum upper frequency in terms of gain. Generally, they are good from audio frequencies up to that limit for amateur experiments.

The maximum safe current of a transistor is important to us also. This is specified as I_c (collector current) for bipolar transistors, and as I_d (drain current) for FETs. At no time should we allow the transistor to draw more current than the specified safe value. In fact, it's wise to operate the device somewhat below (25% or more) that maximum value. This will help to prevent failures from excessive heating of the transistor junction.

A good safety rule is to do all initial circuit testing at reduced operating voltage. For a 12-volt circuit we might want to start our testing at 6 or 8 volts until we were certain that there were no wiring errors. If things seem to be working normally, we can increase the supply voltage to 12.

An "Experimenter's Special"

Thus far we've discussed two rather unprofound transmitter circuits. Once we've finished tinkering with them we may want to move ahead to something more specialized in simple circuitry. Fig. 5 shows the circuit of a two-stage, solid-state QRP transmitter that was designed by Wes Hayward, W7ZOI.[1] Some modifications have been made for this article, but the circuit is essentially as he designed it. This experiment should give us hours, weeks or even months of fun in the workshop and on the air. It delivers slightly more than 1 watt of output to a 50-ohm antenna, and can be made to operate on any band from 160 to 10 meters by using the parts values specified in Table 1. Actually, this is a three-transistor circuit if we count the keying transistor, Q3. But, there are so few parts to the circuit that we can assemble it in short order.

Q1 is a tuned-collector crystal oscillator. Its output energy is fed to the base of Q2, which operates as a Class C amplifier. A pi network (C3, L3 and C4) serves as a harmonic filter (low pass) rather than as an impedance-transformation network, as is more often the case with tube and transistor output amplifiers.

Q3 functions as an electronic switch. When its base resistor is grounded by the cw key it conducts and allows the dc to reach the amplifier stage, Q2. This method helps to reduce the possibility of shorting out the 12-volt supply accidentally, as could happen with the circuits of Fig. 1 and 3 where J1 is in the 12-volt line.

Fundamental crystals are used on 160, 80, 40 and 20 meters. For operation on 15 and 10 meters we will need to use third-overtone crystals at Y1. The oscillator is permitted to run continuously, and keying is applied only to the amplifier, Q2. This prevents chirp on 15 and 10 meters, which would occur if the oscillator stage were keyed.

Feedback capacitor C5 is used only on 160 and 80 meters. All of the component values are the same for 10 and 15 meters. Oscillator trimmer C1 has ample range to provide resonance on both bands.

Construction Thoughts

Experimentation can continue after the transmitter is built and tested — we may want to try our skills at cabinet making, or the unit can be enclosed in a small commercial case, such as one finds at Radio Shack stores. But we can use pieces of double- or single-sided circuit board to fashion a homemade cabinet. We can flow a continuous bead of solder (darned expensive stuff these days!) along the inside seams (corners) of the box to join the side and bottom walls. The lid can be a U-shaped piece of metal (furnace ducting of aluminum). Spray paint or contact paper may be applied to the outer surfaces

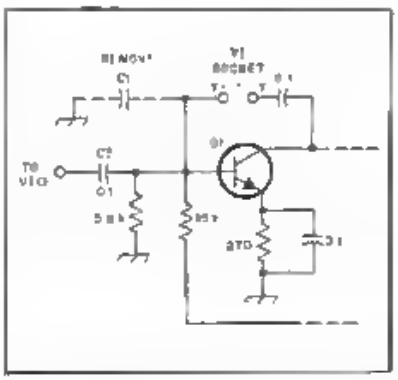


Fig. 4 — Method for attaching a VFO to the circuits of Fig. 1 and Fig. 3. Q1 is thus changed from an oscillator to an amplifier.

Table 1

Fig. 5 Circuit Component Values for Various Bands

	C1 (pF)	C2 (pF)	C3 (pF)	C4 (pF)	C5 (pF)	L1	L2	L3	R1	RFC1
160 m	400	1800	1800	1800	360	73 t No 28 T50-2	8 t	30 t No 26 T50-2	18 Ω	30 t No 28 FT-37-61 (50 µH
80 m	400	100	750	750	200	43 t No 26 T50-2	5 t	21 t No 22 T50-2	39 Ω	21 t No 28 FT-37-61 (25 µH
40 m	180	100	470	470	—	25 t No 26 T50-2	4 t	14 t No 22 T50-2	39 Ω	30 t No 28 FT-37-63 (15 µH
20 m	80	33	210	210	—	27 t No 24 T50-6	3 t	12 t No 22 T50-6	47 Ω	30 t No 28 FT-37-63 (15 µH
15/10 m	80	33	105	130	—	17 t No 24 T50-6	3 t	9 t No 22 T50-6	47 Ω	30 t No 28 FT-37-63 (15 µH

Toroid cores are used in L1, L2 and L3. These are powdered-iron cores available from Amidon Associates and Palomar Engineers (T50-2, etc.). RFC1 is wound on a small ferrite core (FT-37-67), and so on), available from same suppliers. The letter "t" signifies the number of wire turns in the winding.

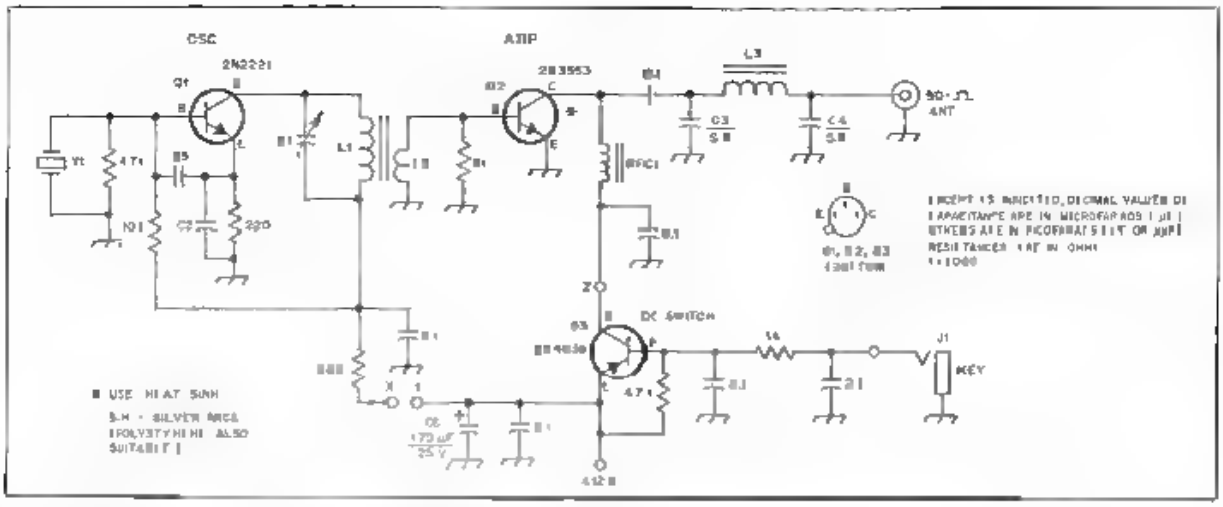


Fig. 5 — Circuit of the W7ZOI "Universal QRP Transmitter." It can provide up to 1.5 watts of rf output when using a 12- to 14-volt dc supply. Fixed-value capacitors are disc ceramic unless otherwise indicated. Resistors are 1/4- or 1/2-watt composition, 10% tolerance. Values not given are listed in Table I. C8 is electrolytic or tantalum. C1 is a mica trimmer. Q2 is a Motorola transistor, but other brands and numbers with equivalent characteristics can be used.

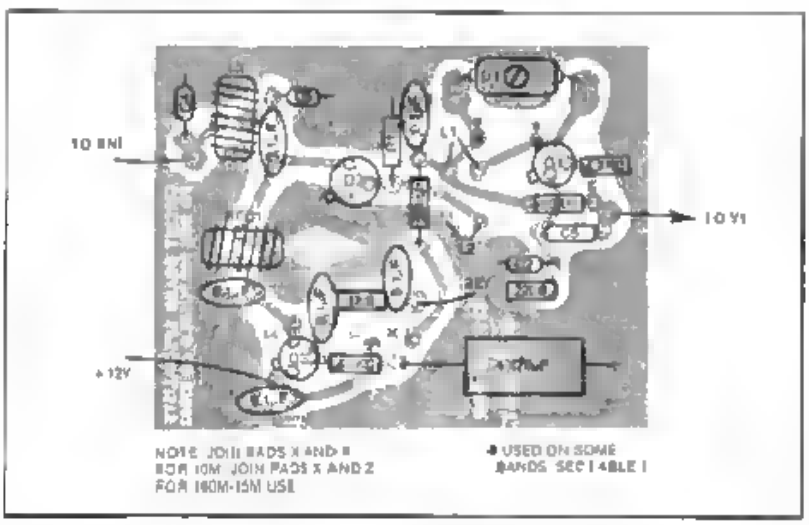


Fig. 6 — Parts-placement guide for the circuit of Fig. 5. The shaded areas represent an X-ray view of the etched side of the board.

of the box to impart that professional look some of us prefer. Press-on decals are excellent for labeling the controls, but Dymo tape labels are suitable also, especially if they are the same color as the panel.

The circuit of Fig. 5 can be assembled on a sheet of pc board using the type of point-to-point wiring described in an earlier *QST* article[1] if a "masterpiece" is not essential to our purpose. But, if pc-board construction of the classic style is preferred we can duplicate the pattern shown in Fig. 6 and in the Hints & Kinks section of this issue.[2] If point-to-point breadboard assembly is our choice we must be careful to keep the input and output components of amplifier Q2 (Fig. 5) separated from one another. Straight-line wiring (not bunched up) is preferable to achieve this. Too-close spacing can cause unwanted feedback and amplifier instability. All of the rf leads in the circuit need to be kept as short and direct as possible. This is especially important when installing the bypass and coupling capacitors.

Caution: When applying operating voltage to the circuits in this article, *check the polarity!* There is no more effective way to send our transistors and electrolytic capacitors on a permanent leave of absence than cross-polarizing the dc voltage connections! Once you have the misfortune of becoming a member of "Junction Busters, Amalgamated," you'll never repeat your mistake!

A Word About QRP Operation

The 1-watt transmitter of Fig. 5 will be 20 dB weaker in signal strength than your transceiver that delivers 100 watts of output. So if you would be heard at 30 dB over S9 with your 100 watts, you will be only 10 dB over S9 with the QRP rig. Or assume your bigger rig was being heard S9 by the other operator. When you switched to the QRP transmitter your signal would drop to roughly S5 or S5-1/2, depending on the accuracy of the S meter (assuming 6 dB per S unit). So you could still be heard well enough under quiet band conditions to be copied "QSL."

Patience and tenacity are the better virtues we can adopt when running low power. Find clear frequencies on which to call CQ. Don't expect answers from stations with weak or marginal signals, unless they are also using QRP. Unless you're a superb operator, it's unlikely that you'll fare very well in DX pileups.

Good antennas are important in successful QRP work. Many first-time QRPers capitulate after a few days of poor results when using mediocre antennas. Erect the antenna high and in the clear, and use a directional, gain type of antenna (beam) on 20, 15 and 10 meters, if you have one available. A good antenna will help to make up for the deficiency in power when using QRP equipment.

The ARRL would welcome clear photographs and reports of the best DX worked with the circuits of Fig. 1 and Fig. 3. Perhaps if we can get enough input on this subject we can run a page of photos, calls and DX records in an issue of *QST*. We

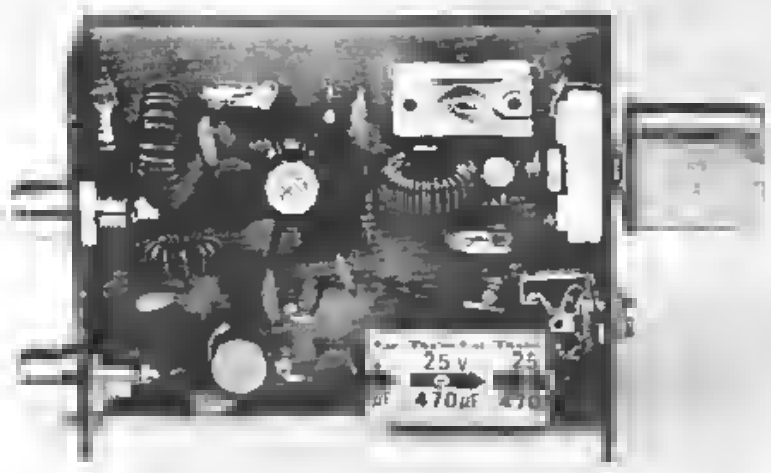
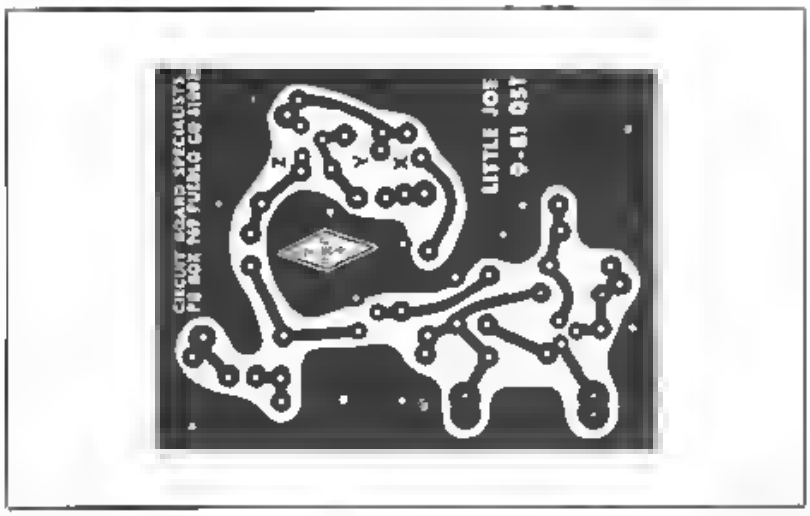


Fig. 7 — Photograph of the assembled kit version (note 7) of the W7ZOI QRP transmitter, as laid out and built by WA6UZO. The panels are made from pieces of double-sided pc board. The dimensions (HWD) are 7/8 × 2 1/4 × 3 inches (22 × 57 × 76 mm).



Etching pattern for the Universal QRP Transmitter. Black represents copper. The pattern is shown actual size from the foil side of the board.

hope you will soon be able to say, "I've built my first piece of amateur gear, and it works great!"

Notes

[1] F. Sutter, "The Rust Sixty" and the 'QSL Sixty,'" *QST*, Sept. 1979, p. 50.

[2] The expression "breadboard" has confused some newcomers to Amateur Radio. It originated in the early days of the amateur service when home-built transmitters on wooden foundations, such as the ends from orange crates. The kitchen breadboard became popular for that purpose, and thereafter any wooden chassis base was called a breadboard.

[3] Details for building a simple diode rf probe can be found in the measurements chapter of the past several editions of *The Radio Amateur's Handbook*.

[4] D. DeMaw and R. Shriver, "A Simple Utility Power Supply," *QST*, Nov. 1979. Parts kits available from supplier in note 7.

[5] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur*, Newington, CT: American Radio Relay League, Inc., 1977), ch. 2, p. 36. This publication is recommended for experimenters because it contains a wealth of basic theory and many practical examples of simple transmitters, receivers and test equipment.

[6] D. DeMaw, "Quick and Easy Circuit Boards for the Beginner," *QST*, Sept. 1979, p. 30.

[7] Etched and drilled circuit boards for the transmitter are available from Circuit Board Specialists.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

From September 1979 QST, p 30:

Quick-and-Easy Circuit Boards for the Beginner

Why endure layout agony and the mess of chemicals? Make your own breadboard-style modules quickly. Here's how!

By Doug DeMaw, W1FB
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Luther, MI 49656

If you dread those brown ferric-chloride stains on your clothes, the tedium of etching a circuit board and the puzzlement of laying out a pc-board pattern, this article is for you! There's no rule that suggests a firm need for commercial-quality circuit boards. Sure, the professional stuff looks great with those lines and circles of copper so neatly etched on epoxy or phenolic board material. But, consider the person hours involved in planning a layout, applying the etch-resist tape or lacquer, then etching away the unwanted copper. All of this can become rather futile if the amateur is interested only in testing a circuit on a one-shot basis. A simple breadboard type of assembly will often suffice; time and money will be saved in the process.

But what of the finished product? Sure, nobody really wants an "ugly duckling" to show off at the next club meeting or when hams drop in to visit the shack. However, good looking circuit-board assemblies can be had even when using the non-etch techniques outlined here. An experimental circuit board which has the components neatly in place, "dress-right-dress" fashion, can be a thing of beauty to the beholder's eyes, provided he or she is not an inspector for a government-contract job! Let's examine some ways to make our own non-etched boards.

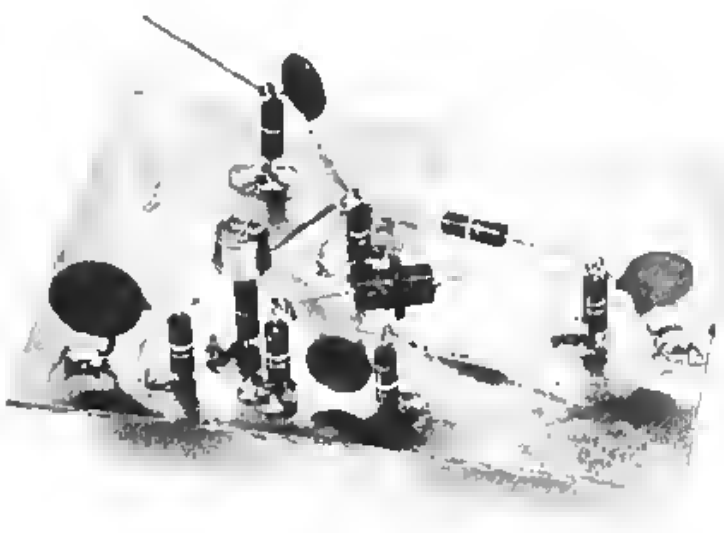


Fig. 1 — Photograph of the simple breadboard depicted in Fig 2A. High-chmega resistors serve as standoff terminals.

The Standoff Technique

The basic foundation for any of the "quickie" boards we shall discuss here is a sheet of copper-clad circuit-board material — the kind we find in proliferation at hamfest flea markets, Radio Shack stores and similar outlets. It need not be clad on both sides, but "double-sided board," as the near misnomer indicates, is suitable also. Our objective in making any circuit board ready for use is to provide a suitable number of electrically isolated conductive islands upon which the various components can be connected by means of solder. At least that is the fundamental principle of *etched* boards. But, alternative methods exist for developing isolated pads or tie points. Regardless of the approach taken to achieve this effect, the name of the game remains the same. Assemble the components close to and above a copper "ground plane." In this type of situation the copper on the board becomes the circuit ground, just as a chassis does when circuit boards aren't employed. The copper plane enables us to make direct ground connections, thereby minimizing the pigtail lengths of the various resistors, capacitors and transistors. Short lead lengths and a quality ground conductor help prevent circuit instability (self-oscillations or parasitics). It is for this reason that many amateur projects call for double-sided pc board. One side is etched and the opposite side is solid copper, except where the various

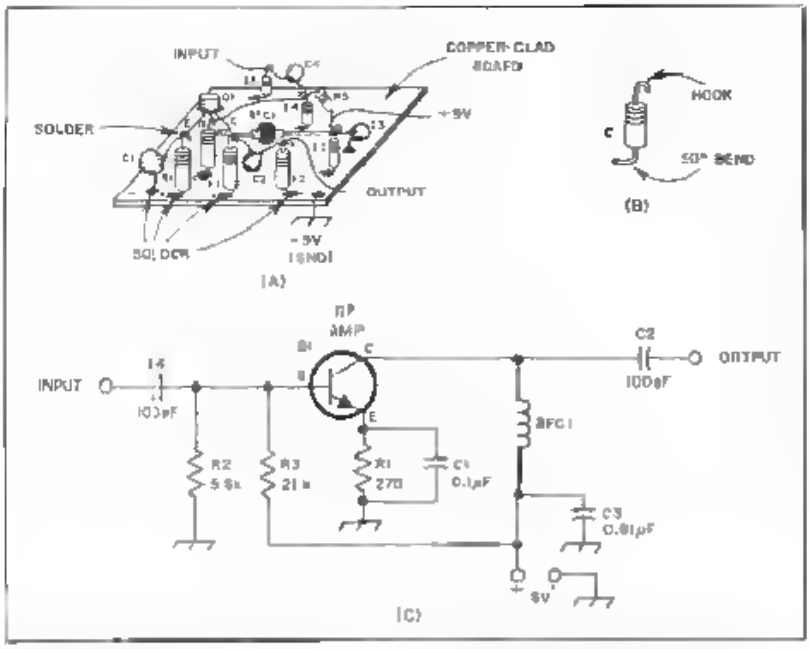
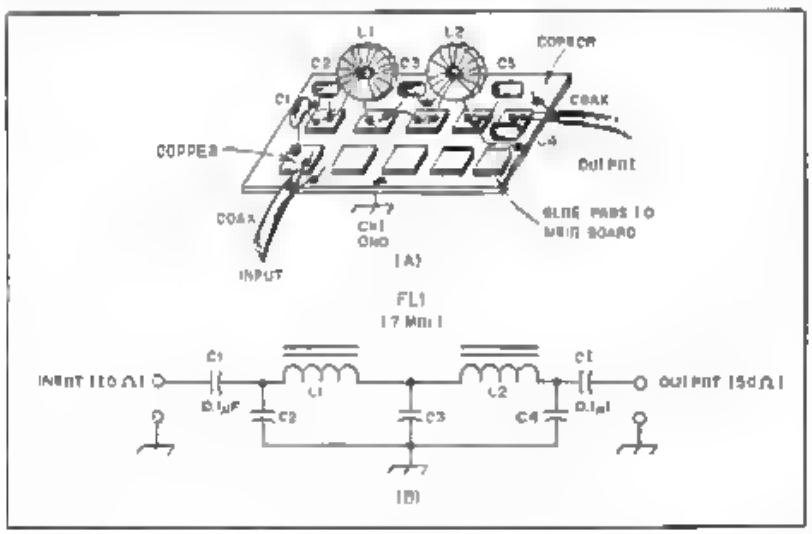


Fig. 2 — Pictorial view of a circuit board that uses the standoff technique as shown at A. A closeup view of one of the standoff resistors is shown at B. Note how the pigtails are bent. The schematic diagram at C shows the assembled circuit displayed at A.

Fig. 3 — Illustration A shows how isolated pads can be glued to a piece of copper-clad pc board to form a simple breadboard. The equivalent circuit is shown at B. A short wire jumper can be used to join individual pads as shown just below C3 at A.



components are installed. A small amount of copper is removed around each pc board hole to prevent short-circuiting the component leads to the ground plane.

The "standoff technique" calls for some more haggling at flea markets. Along with the pc-board material we gamed in the swap-and-shop area, we must look for high-ohmage 1/4- or 1/2-watt composition resistors. Values from 220 kΩ to several megohms are suitable. Generally, bags of bargain resistors are abundant. Don't worry if the pigtails are short; this type will be just fine, provided the wire leads are each 1/4 inch (6 mm) long or greater.

The purpose of utilizing high-ohmage resistors is to ensure that they act more like *insulators* than as true resistors. Hence, the higher the resistance the better for our application. As a rule of thumb, the resistor being used as a standoff insulator should have a value that is at least 10 times the circuit impedance or value of resistance used at that circuit point. For example, if a resistor is used as a tie point at the 50-Ω output of a circuit, the resistor should be a 500-Ω type, or greater. In the circuit at Fig. 2C we find that R2 is 5.6 kΩ and R3 is 27 kΩ. We will make the standoff resistor, E4, 10 times greater than the smallest resistor, R2, or 56 kΩ. Any value higher than 56 kΩ will work nicely, too. Our only concern for the value of E3, the +9-volt standoff resistor, is that it is high enough in value to minimize the current drain from our power supply. A value of 220 kΩ would draw only 0.04 mA at 9 volts. This is an insignificant amount, even if a transistor radio battery served as the power source.

Fig. 2A illustrates pictorially how one might apply the standoff technique to wire the circuit shown at C of Fig. 2. Illustration B demonstrates how the resistor pigtails are bent before the component is soldered to the pc-board material. The base circuitry at Q1 of Fig. 2A is strung out excessively. This was done to reduce clutter in the drawing. In a practical circuit all of the signal leads should be kept as short as possible. E4 would, therefore, be placed much closer to Q1 than the drawing indicates.

R1 and R2 of Fig. 2 actually serve two purposes: They are not only the normal circuit resistances, but function as standoff posts as well. This practice should be followed whenever a capacitor or resistor can be employed in the dual role.

Glue-and-Pad Method

A simple but more time-consuming technique for making circuit boards is illustrated in Fig. 3A. Once again we have chosen a piece of copper-clad pc board as our foundation. The isolated pads consist of small squares of pc-board material, single or double sided. The size of the pads is arbitrary, but the author prefers them to be 1/4 inch (6.3 mm) square. The smaller the pads the more of them can be placed in a given area of the main board. There is no reason the pads can't be rectangular or round. The format will depend on personal choice and the cutting technique available to the amateur.

A power or "armstrong" type of hack saw is perhaps the best tool for cutting the board material into squares. If you have a friend who has access to a model-shop sheet, perhaps he will be willing to cut a supply of squares for you during his lunch hour. Phenolic-based pc board will shatter in a sheet, but glass-epoxy board with a cloth base will cut cleanly when sheared.

Once the squares are prepared they can be affixed to the main board by means of epoxy cement. It can take as long as 24 hours for the glue to set firmly, depending on the brand and grade of cement used.

Noted *QST* author W7ZOI once suggested an alternative method for gluing pads to a pc board. He mentioned household glue as a faster agent for attaching

the pads to the main board. This type of adhesive is available from hardware stores in small packets. It comes in tubular stick form. A thin slice of the hard glue is shaved from the stick by means of a knife. The glue slice is placed between the bottom of the pad and the top surface of the main board. A soldering-iron tip is pressed against the pad, held there until the glue melts and spreads, then removed.

The shortcoming of this approach is that the glue softens each time a component is soldered to a pad. It requires more than casual care when assembling the circuit components. Epoxy glue will not melt during the soldering process. Also, the completed module will remain intact much longer if epoxy cement is used.

A comparison between the practical and schematic circuits for a simple half-wave harmonic filter is provided in Fig. 3. The pictorial version shows how we might mount the parts when using the glue-and-pad technique. There are some unused pads in the foreground.

Saw-Slot Boards

For those who subscribe to the Armstrong method discussed earlier, a hack saw can be used to cut through the copper on a section of pc board to form isolated pads. This concept was popularized by the author and W1ICP in a *QST* beginner's series which ran from April through September 1974. An example of this technique is given in Fig. 4. The copper must be cut *completely* away where each dark line is shown. This will prevent short circuits between adjacent pads. A hobby Moto Tool can be employed to cut the grid seen in Fig. 4. If this is done, a straight-edge guide will be necessary if a neat job is desired.

If Etching Is Your "Thing"

A universal pc breadboard can be etched and used many times if one does not mind dabbling with etchant chemicals. A few brown stains here and there on one's garments could fit nicely into our contemporary world of fashion, so maybe the etching technique isn't all that bad!

The pattern shown in Fig. 5 is arbitrary. It suits the author's needs handily for circuit experimentation. The smaller pads are suitable for mounting transistors, diodes and other discrete components. The long conductors along the edges and one side of the breadboard are laid out to function as plus-voltage and ground buses. The +V foil has branches that extend through four groups of pads. The ground conductor has similar branches that pass through alternate groups of pads. This helps to keep lead lengths on the various parts to a minimum.

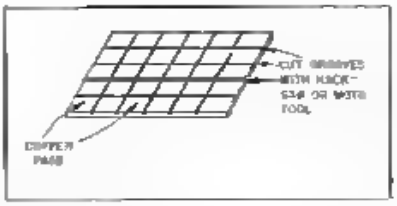
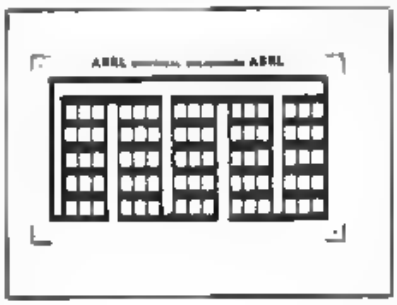


Fig. 4 — Saw-blade scoring of copper-clad pc board material provides isolated pads in this drawing.

Fig. 5 — Suggested pattern for a universal breadboard that can be etched as shown, not to scale. A +V and a ground bus are provided for easy access along the length of the board. See the "Hints and Kinks" section of this issue for a full size pattern template.



We can use ordinary masking tape as the etch-resist material. Cover all of the copper surface with a layer of tape. Press the tape firmly against the copper by running a smooth object, such as the side of a lead pencil, back and forth across the tape. Next, draw the pattern on the tape. An X-acto knife can be utilized to remove tape sections where the copper must be etched away. All that remains for board preparation is 15 to 30 minutes of etching in ferric-chloride solution. *Do not allow the etching solution to contact your eyes or skin. If it does, wash it off immediately with clear, cool water.*

To ensure reasonable longevity of this type of breadboard, material with heavy-gauge copper laminate should be used. Glass-epoxy insulation is also recommended. If the copper is too thin, repeated soldering will loosen the copper, and stress on the pads will separate them from the main board.

Some Closing Comments

Needless to say, earlier comments about brown stains on one's clothing were offered in a purely jocular vein. Avoid allowing the etching chemicals to splash on your clothing! The stain will be permanent!

There are probably a number of additional methods for fabricating circuit boards quickly and simply. No credit is claimed for originality concerning the procedures described in this article. The purpose of this presentation is to illustrate some of the more common approaches to breadboard fabrication without chemicals. But of greater importance, we've tried to stimulate confidence among those who were heretofore unwilling to engage in home-project work through fear of circuit-board layout and etching. Let's compare brown stains later if we should use ferric chloride. If not, perhaps we can swap high-value resistors or sharpen saw blades together!

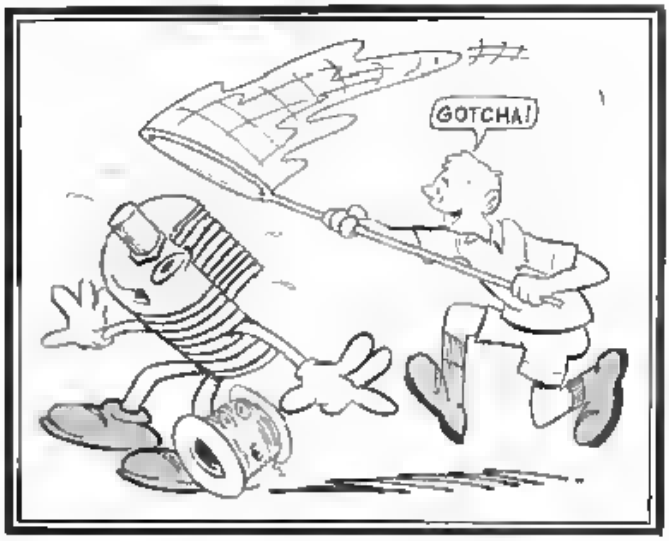
Black represents copper; the pattern is shown at actual size. The board is single sided (copper on one side only), shown from the foil side, and is a universal breadboard pattern suggested by DeMaw (see Fig. 5).

[1]Negatives and circuit boards for the breadboard are available from Circuit Board Specialists.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

From October 1987 QST, p 24:

Stalking Those Fugitive Components



Specialty components appear hard to find for those who aren't experienced gleaners. Let's learn where and how to obtain some of these bread-and-butter items.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther, MI 49656

What's this you're saying? You would build more ham gear if only you could obtain the necessary components? I receive dozens of letters to this effect each year. Most of them seem to be from the newer hams who have yet to learn the fine art of foraging for those seemingly elusive parts. Some correspondents are critical because my *QST* articles are not based on using parts that can be purchased at Radio Shack stores. Sure, Radio Shack stocks a lot of things that are useful for building projects, but many of the circuits we amateurs want to build require components that Radio Shack will never carry. A designer is severely restricted if he has to rely on any single supply source. At best, his output will soon be reduced to kinky-dink projects.

What, then, might you do to solve the annoying parts-procurement problem? This subject has been addressed frequently in *QST*, but only in general terms. That is, the authors did not focus on specialty items that many of us need from day to day. This article is aimed at those unique parts that we do not find at the corner parts store. All you need is some ambition and a few postage stamps to equip yourself with the means to get the parts highlighted here.

Some of the suppliers I list in this article have many parts to offer in addition to those discussed here, and numerous other suppliers exist. I concentrate in this article on those dealers from whom I purchase most of my parts and materials. I consider their prices fair and generally below the figures set by new parts distributors that aren't in the surplus business. I have experienced neither poor service nor rip-offs from any of the dealers listed, but neither the ARRL nor I endorse them. As the saying goes, "let the buyer beware."

Locating Component Sources

I watch for some of the smaller display ads in *QST* and other amateur publications, and keep tabs on the classified ads in the various magazines. That is where you'll often see information that can lead to a free catalog of bargain parts. I respond to every ad of that type. Consequently, I have stacks of catalogs. It is a practice I recommend to all of you who enjoy building amateur equipment. There is scarcely a component I can't find for my projects, if I scan the pages of these mail-order catalogs.

Writers (myself included) often recommend ham-radio flea markets as a source of parts for home use. Flea markets are, indeed, wonderful places to look for certain items. But, owing to the infrequency of flea-market events in any given region, procuring parts by that means is a long-range situation at best. I depend on flea markets mainly to stock up on items for future, unplanned projects. For example, if I see a super bargain on 2N2222s, polystyrene capacitors or 2200-µF filter capacitors, I buy them for later use. This practice also enables me to help other hams in the area, should they have a sudden need for something I have in my goodie cache.

Parts and materials never appear magically! We may daydream until doomsday, but that won't yield results. We must also innovate as the demand dictates.

Equipment Cases

Consider the low cost and simplicity, for example, of fashioning a small project case from galvanized furnace-ducting material. Most plumbing and heating shops will give you scraps or pieces from stock, or they may charge you a few cents per pound for the material. A large pair of tin shears can be used to cut the sections of metal to shape, and bending can be done by hand over any right-angle form. The cabinet walls and top can be soldered together, or fastened with no. 6 sheet metal screws. The completed cabinet can be spray painted with sandable gray primer, sanded and then coated with your favorite color of paint for the finishing touch.

Large cabinets, such as those used for antenna-matching networks, can be fashioned from tempered Masonite®. This material can be painted any color you prefer. The front panel can be made from an aluminum cookie sheet, available at most variety stores. There is no need to contain a Transmatch in a shielded cabinet, since it does not generate TVI. The signal going into the Transmatch should already be clean!

I have mentioned many times the ease and low cost of making small boxes from sections of single- or double-sided PC board. The cost of any of these homemade enclosures is substantially less than that of a commercially made box, and the materials are available locally. These methods permit almost instant construction of an equipment case.

Magnet Wire

Many hams ask me where they can find magnet wire. I must say that the market has, for the most part, dried up with respect

to magnet wire. Radio Shack sells small spools of enameled wire, but only in a few popular gauges. Jug Wire Co in New York was my primary source for magnet and bare bus wire, but a recent notice from Jug indicated that they were going out of business.

What can you do to solve this problem? First, check with your local electric-motor repair shops. The operators are often willing to reel off a reasonable number of feet of the wire you need, and at a nominal cost. Here, again, use your imagination.

When I first became a ham, it was common practice for my colleagues and me to acquire old power transformers just for the purpose of removing the magnet wire from the windings. The same was true for old dynamic speakers from junked radios. The speaker field coils contain hundreds of feet of small enameled copper wire! Still another source of magnet wire is the field coils of large, low-resistance dc relays—12- and 28-V units in particular. Generally, the larger the relay and the lower the field-coil resistance, the larger the wire gauge. Look for these relays at flea markets. They can be available for 25 cents or less.

Another excellent source of magnet wire is picture-tube yokes from discarded TV receivers. The vertical and horizontal deflection coils contain many feet of usable sizes of wire.

Litz Wire

Litz (short for *litzendraht*, which means "stranded wire") wire is desirable for winding small LF, MF and HF slug-tuned coils. It provides a higher Q than plain enameled wire. This is because many strands of enameled wire are used to form a cotton- or silk-covered conductor. The additional surface area afforded by multiple conductors offsets *skin effect*—the tendency for ac to flow at or near the surface of a conductor, resulting in greater ac resistance with rising frequency. I have never seen Litz wire offered in surplus equipment catalogs. I obtain my Litz wire by purchasing old RF chokes and slug-tuned coils that are wound with it. Many WW II power RF chokes contain Litz wire, and you may want to consider this method of garnering some.

Coil Forms and Insulating Material

Blank slug-tuned coil forms are currently too expensive to consider for most amateur projects. There are some surplus bargains, however, and you should watch for them. Stock up on these forms should you see them at flea markets, but be aware of the effects of improper core material on operating frequency. Low-frequency cores will spoil the Q of an HF or VHF slug-tuned coil. The same is true of improper toroid-core material. A relative test of coil Q may be made by winding a coil on an unknown form, then placing a silver mica or variable capacitor in parallel with the coil to obtain resonance at a desired frequency. Check the tuned circuit with a dip meter. If a good dip can be had with the dipper coil a fair distance from the test coil (say, one inch), the Q is reasonably high. If, however, the dipper must be coupled tightly to the test coil to obtain a dip (usually shallow at best), the Q is probably too low to consider for your circuit. In other words, the farther the dipper coil is from the test coil, consistent with a deep dip reading, the higher the Q of the coil.

Homemade fixed-inductance coils can be wound on plastic tubing and rods as a low cost. Included are power-line RF chokes, antenna traps and antenna-loading coils. First, check the scrap department of your local plastics outlet for odd-lot bargains. Such materials as Teflon®, Delrin, polystyrene, Plexiglas™, PVC, Tenite® and Lexan® are often available in small pieces at attractive prices. You may also obtain a catalog from United States Plastic Corp in Lima, Ohio, a mail-order house that has tubing, rod and sheeting of all types (see listing at the end of this article).

Feed-line spreaders can be made inexpensively from such materials as hair curlers, plastic clothespins, sections of plastic coat hangers and even ballpoint pen bodies. Again, I stress the value of being innovative!

Special Capacitors

I've read many laments about how "impossible" it is for some *QST* readers to locate high-voltage disc-ceramic capacitors, polystyrene capacitors, NP0 capacitors and even silver-mica capacitors. These items are *widely* available from the surplus-parts vendors. I must admit, however, that large transmitting variable capacitors *are* scarce (and extremely expensive). Radiokit seems to be the main outlet for large variable capacitors. When the Cardwell Corp bought the tooling and stock of E. F. Johnson and Hammarlund several years ago, it seemed that a variable capacitor monopoly was taking shape. The James Millen Co was the only other major manufacturer of these parts and, to complicate matters more, Millen went out of business, too. It was a sad day for Amateur Radio! You may still be able to obtain Millen capacitors from Radiokit. Our best hope is to remove large variable capacitors from surplus radio gear, such as WW II command transmitters and BC-191/BC-375E transmatch tuning units. Fair Radio Sales in Lima, Ohio is worth checking for these units and other large WW II electronics equipment. Their catalog will fill many of you older hams with nostalgia!

The Joys of Stripping

Lest someone misunderstand, I refer to radio parts! When Lew McCoy, W1CF, was the Beginner and Novice editor for *QST*, years ago, he constantly stressed the value of stripping parts from old TV and radio sets for use in ham projects. I'm sure that many of you recall his "Transmitters from old TV sets." We at ARRL HQ often wondered why he never made a TV set from *an old transmitter*, but he refused the challenge when it was offered to him! Nonetheless, his advice in those days was sage. Even today we can glean countless excellent small parts from old TV and transistor-radio sets. I saw six table-model TV sets for sale last fall at the Hudsonville, Michigan, ham flea market. The owner was asking 25 cents apiece for the sets! Many PC-mount fixed and slug-tuned coils are found in TV receivers, in addition to a host of resistors and capacitors. Also, you can

salvage many potentiometers and switches, as well as a variety of hardware to add to your stock of nuts and bolts.

Pocket-size transistor radios are loaded with small resistors and capacitors. How many of these little radios have you thrown away when they became defective? Consider the parts you could have salvaged for later use. Discarded AM and FM receivers also contain small variable capacitors that can be used for homemade receivers and QRP transmitters. The IF transformers can be used as is, or can be rewound for other frequencies. Not only can you increase the bulk of your parts larder by stripping TV sets and transistor radios, you will have a nice pastime for those rainy or snowy evenings in winter. Solder wick or solder suckers are invaluable for this job.

Source Listing

Table 1 lists a number of hard-to-find components keyed to the suppliers that stock them. The dealer identification is given at the bottom of the table. I have identified specific components that are offered by these suppliers, but they carry many additional items. Their catalogs are worth adding to your reference library. Remember that quantities and specific values may be limited, depending on the supplier.

Some Final Comments

Although this month we haven't covered theory, applications or a practical project, I feel that parts procurement is an important part of construction. I have addressed those parts that readers seem to have the greatest difficulty locating. Perhaps this article will reduce the number of inquiries I receive!

Unfortunately for us amateurs, some of the suppliers listed specify a minimum order. In such instances, it is sometimes convenient to pool your order with those of other hams in your area. This may require some salesmanship on your part, but it can be done. Good luck in stalking those fugitive components!

ARRL Parts Suppliers List

S E J L M W Y
A&A Engineering
2521 W LaPalma Ave Unit K
Anaheim, CA 92801
714-952-2114
*sase

A B C F G H I J K L M N S X Y
All Electronics Corp
15004 Oxnard Street
Van Nuys, CA 91411
800-826-5432
*free **$10

A C E I K L S X
Allied Electronics
401 E 8th St
Fort Worth, TX 76102
800-433-5700
**$25

A B C D E G H I J K L M U X Y
Alpha Electronic Laboratories
705 Vandiver Suite A
Columbia, MO 65202-2052
314-874-1514
*free

B
Amidon Associates Inc
12033 Otsego St
N. Hollywood, CA 91607

A D F G H Miller components
Antennas Etc Unlimited
PO Box 4215
Andover, MA 01810-4215
508-475-7931
*sase

A E F G I K L S X
Arrow Electronics
25 Hub Dr
Melville, NY 11747-3928
800-932-7769 fax 516-694-6870
**$25

N O facsimile equipment
Atlantic Surplus Sales
3730 Nostrand Ave
Brooklyn, NY 11224
718-372-0348
*$1

I K M T U
ATV Research Inc.
13th & Broadway
Dekalb City, NE 68731
402 987 3771

Avatar Magnetics
see Ronald C. Williams

A,D,H
Barker & Williamson
10 Canal St
Bristol, PA 19007
215-788-5581

A,B E G,H I,J,K,L,M,N,U,X,Y
BCD Electro
PO Box 450207
Garland, TX 75045-0207
214-343-1770 fax 214-343-1854
*$1/year

A,B C,D,E,G,H,I,W,
Circuit Boards for ARRL projects. Info
Circuit Board Specialists
PO Box 951
Pueblo, CO 81002-0951
719-542-4625
*free

A,B,C,E,I,K
Circuit Specialists, Inc.
PO Box 3047
Scottsdale, AZ 85271-3047
602-966-0764
*$1 **IE (mail orders)

A B D E H I T W
Communications Concepts, Inc.
508 Millstone Drive
Xenia, OH 45385
513-426-8600 fax 513-426-3811
*free

L,E tools
Contact East Inc
335 Willow Street
N Andover, MA 01845
508-682 2000
*free

Bowers
Coolironic
1700 Mesa Ave
Ventura, CA 93003
805-642-8521
**$50

D,E,I,M H X
Peter W. Dahl Co. Inc
5869 Waycross
El Paso, TX 79924
915-751-2300

F
Davis RF
PO Box 230
Carlisle, MA 01741
508-369-1738
*$1

A,D,E,F,H,I,K,L,N,S,W,X
Dick Smith Electronics
PO Box 468
Greenwood, IN 46142
317-888-7265

A,E,I,J,K,L,X Y
Digi-Key Corporation
701 Brooks Ave. So.
PO Box 677
Thief River Falls, MN 56701
800-344-4539
*free

C
E-Z CIRCUIT by Bishop Graphics, Inc
20938 Roscoe Blvd.
Canoga Park, CA 91304
818-773-9081
*free **$25

A,C,E,F,I,K,L,N,S,X
Edw Electronics, Inc
2700 Hempstead Tpke
Levittown, NY 11756-1143
516-735-3330

A,C D E,F,G,H I,K,L,S,X,Y
Electro Sonic, Inc
1160 Gordon Baker Rd
Willowdale, Ontario Canada M2H 3B9

A,L
Electronic Distributers, Inc
4900 N. Elston Ave
Chicago, IL 60630
312-283-4800

B
Fair Radio Sales Co., Inc
PO Box 1105
1016 E. Eureka St.
Lima, OH 45802
419-227-6573 fax 419-227-1113
*free **$10

W
FAR Circuits
18N640 Field Ct
Dundee, IL 60118
708-426-2431

J (8-pole crystal filters)
Fox-Tango Corp
747 S Macedo Blvd
Port St Lucie, FL 34983
407-879-6868
*sase

D
Gregory Electronics
249 Route 46
Saddle Brook, NJ 07662
201-489-8000
free 525

A B I K M N T
H & R Corporation
A. Hirbich & Rademan Co
401 E. Erie Ave
Philadelphia, PA 19134
215-426-1700
*free **$10

I,K
Hammond Mfg. Co., Inc
1630 Walden Ave.
Buffalo, NY 14225

I,K
Hammond Mfg. Ltd
394 Edinburg Rd. N
Guelph, Ontario Canada N1H 1E5

D
HiMouldh
PO Box 802
Council Bluffs, IA 51502
*$1 Catalog req'd

A B E H I K L
Hosfelt Electronics
2700 Sunset Blvd
Steubenville, OH 43952
800-524-6464

B J L O W (8 pole crystal filters)
Int'l Radio and Computers Inc
751 E Macedo Blvd
Port St Lucie, FL 34983
407-878-5885
*sase

Y
International Crystal Mfg Co.
PO Box 26330
Oklahoma City, OK 73126
405-236-3741

A,D,F G H,
Miller Equipment
James Miller Electronics
PO Box 421SBV
Andover, MA 01810
508-975-2711 fax 508-474-8919

Y
JAN Crystals
2341 Crystal Dr
PO Box 06017
Ft. Myers, FL 33906-6017
800-526-9825
*free

E
K2AW's Silicon Alley
175 Friends Lane
Westbury, NY 11590
516-334 7024
*free **$10

C
Kepro Circuit Systems Inc
630 Axminster Dr
Fenton MO 63026-2992
800-325-3878 (out of state)
314-343- 630 (MO)
*free **$15

C,E,F,G,I,K,L,X
Lashen Electronics, Inc.
21 Broadway
Denville, NJ 07834
201-627-3763 **$15

A,L, EXAR XR-2208 IC
Marshall Electronics
20 Sterling Dr
Wallingford CT 06492
203-265-3822
**$25

F (aluminum and stainless tubing and pipe
Metal and Cable Corp Inc
PO Box 117
Twinsburg OH 44087
216-426-8455
**$50

K,L
MFJ Enterprises
PO Box 494
Mississippi State, MS 39762
601-323-5869

A,B,E,H,M
Microwave Components of Michigan
PO Box 1697
Taylor, MI 48180
313-753-4581 (evenings)
*base

A,B,C,D,E,F,G,H,I,J,K,L,S,U
Milo Associates, Inc
4166 Millersville Rd.
Indianapolis IN 46205
317-546-3456 fax 317-547-1729
*free

mmwave signal processing
Mini Circuits Labs
PO Box 350166
Brooklyn, NY 11235-0003
718-934-4500

A,E,F,G,H,I,L,M,X,Y
Modern Radio Laboratories
PO Box 14902 Q
Minneapolis MN 55414
*$2

F
Mosher Instrument Co
Ivoryton Industrial Park
Main Street
Ivoryton CT 06442
203-767-2104 800-243-8316
**$10

F
Motorola Semiconductor Products, Inc
PO Box 20912
Phoenix, AZ 85036

A,B,C,E,F,G,H,I,J,K,L,S,U,X,Y
Mouser Electronics
2401 Hwy 287 N
Mansfield, TX 76063
800-346-6873
*free **$20

E
National Semiconductor Corp
2900 Semiconductor Dr
Santa Clara, CA 95052-8090

A,D,E
Oak Hills Research
4061 N Douglas Rd
Luther, MI 49656
*base

A,B,C,D,E,F,G,H,I,L,K,L,M,N,S,T,U,W,X,Y
Ocean State Electronics
PO Box 1458
Westerly, RI 02891
800-866-6626 fax 401-596-3590
*free **$5

A,E,J,L
ORA Electronics
9410 Owensmouth Ave
Chatsworth, CA 91311

B
Palomar Engineers
PO Box 455
Escondido, CA 92025
619-747-3343
*free

A,D,X
Pasternack Enterprises
PO Box 16759
Irvine, CA 92713
714-261-1920
*free

T
PC Electronics
2522 S Paxson Lane
Arcadia, CA 91007
818-447-4565 fax 818-447-0489
*free

B
Philips Components (Ferroxcube)
Mr Marlen Rd
Saugerties, NY 12477

A,L
R & D Electronics
1224 Prospect Ave
Cleveland, OH 44115
216-621-1052

A,D,E, Electron tubes and RF power transistors
R.F. Gain, Ltd
116 South Long Beach Rd
Rockville Centre, NY 11570
800-348-5580

A,B,C,D,E,F,G,H,I,J,K,L,S,U,W,X,Y,Z
Radiokit (mail)
PO Box 973
Pelham, NH 03076
Telex 887387
*$1

Radiokit (store)
15 Londonderry Rd Unit 8
Londonderry, NH 03053
603-437 2722

A,B,D,E,F,H
RF Parts Co
1320-10 Grand Ave
San Marcos CA 92069
619-744-0728
*free **$20

D,Y
Sentry Mfg., Co
Crystal Park
Chickasha OK 73018
405-224-6780
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E
Silicon General, Inc
11861 Western Ave
Garden Grove, CA 92641

(Mechanical components and metal stock)
Small Parts Inc.
PO Box 381900
Miami, FL 33238
305-751-0856
*free **$12

A,D,F,J,Y
Spectrum International Inc
PO Box 1064
Concord, MA 01742
508-263-2145

A,M,N
Star-Tronics
PO Box 683
McMinnville, OR 97128
503-472-9716
**$4

M,N
Surplus Traders
PO Box 276
Alburg, VT 05440
*$1

K,H
Ten-Tec, Inc.
Highway 411 E
Sevierville, TN 37862

I
Toroid Corporation of Maryland
6000 Laurel-Bowie Rd.
Bowie, MD 20715-4037
301-464-2100
*free

A,G,O tubes
Typetronics
PO Box 8873
Ft Lauderdale, FL 33310
*base

F
US Plastic Corp
1390 Neubrecht Rd
Lima OH 45801
800-537-9724

I,Z
Ronald C Williams W8JVF (Aveiar Magnetics)
1408 W Edgewood Ave
Indianapolis IN 46217

A,E,F,T,W
Wyman Research, Inc
Box 35 R R 1
Weldron, IN 46182
317-525-6452

Chart Coding

A—New Components
B—Te olds and Fairlies
C—Etched Circuit Board Materials
D—Transmitting and Receiving Materials
E—Solid-State Devices
F—Antenna Hardware
G—Dials and Knobs
H—Variable Capacitors
I—Transformers
J—IF Filters
K—Cabinets and Boxes
L—General Supplies
M—Surplus Parts
N—Surplus Assemblies
O—R**TY Equipment and Parts
P—Surplus FM Gear and Parts
Q—Equipment Manuals
R—Service of Collins Equipment
S—Test Equipment
T—Amateur TV Cameras and Components
U—Microcomputer Peripheral Equipment
V—Tweeks
W—Ready-made Printed Circuit Boards
X—Wire
Y—Crystals
Z—Climbing and Safety Equipment

*Catalog Price
**Minimum Order

To the best of our knowledge the suppliers shown are willing to sell components to amateurs in small quantities by mail. This listing does not necessarily indicate that these firms have the approval of ARRL.

From September 1988 QST, p 44:

BLENDING CIRCUIT-BOARD FABRICATION TECHNIQUES FOR SUCCESS

☐ In his August 1987 article on homemade circuit boards, Doug DeMaw mentioned the unsuitability of mechanically etched boards for use with ICs or other components with close pin spacings. (Generally, mechanical etching isn't precise enough to make traces suitable for the 0.1-inch pin spacing standard with ICs.) I've been getting around this limitation by making a gridded sub-board for the IC and mounting it to the main (mechanically etched) circuit board with the piggyback method described in Doug's article (see Fig 4). Jumper wires connect the IC sub-board pads to the main circuit board; glue holds the IC sub-assembly in place.—*John Evans, K3SQO, RR1, Box 131, Kingsley, PA 18826*

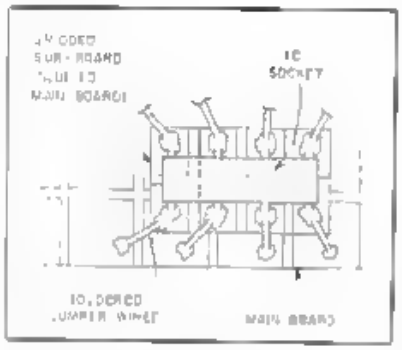


Fig 4—John Evans gets around the incompatibility of mechanically etched boards and ICs by mounting his ICs on gridded, single-pad sub boards. (Hint: the main board is also gridded for clarity.) The sub-boards are mounted to the main board using the piggyback technique described by Doug DeMaw. See text.

ETCH-RESIST PENS FOR HOME-MADE CIRCUIT BOARDS

☐ Because I've been fabricating circuit boards at home for some time, Doug DeMaw's circuit-board article[1] was of more-than-usual interest to me. In particular, I've been involved in "longhand" PC-board production (a general term for boards produced with resist applied by hand with a brush or marking pen) for quite some time.[2] Most problems with boards made by the longhand method are caused by uneven ink flow from the pen. Marcus referred to this problem in a *CQ* article.[3] This uneven-flow problem can be corrected by opening the pen and adding a solvent that is compatible with the ink. (Usually, the ink vehicle is an alcohol-based solvent.)

The ink in most felt- or fiber-tip pens is stored in a fiber cylinder enclosed in a thin plastic sheath. Add 10 to 15 drops of alcohol or a similar solvent [rubbing alcohol [70% isopropyl], lacquer solvent [denatured ethyl alcohol] and butyl acetate [thinner for model paints] are satisfactory] to the cylinder end that contacts the pen tip. (Stop adding alcohol if it appears that the next drop will cause leakage from the bottom of the cylinder.) Replace the ink cylinder in the pen and allow a few minutes for the rejuvenated ink to migrate into the pen tip. Now, the pen should produce opaque black lines without smearing. If the lines appear to be almost *too* fluid, that's ideal. (By the way, overapplication of alcohol to the ink cylinder can cause leakage through the pen's tip vent hole. Watch out for this so you don't generate profanity when a vent drop hits the board and spoils your work!) Using this method, I've successfully rejuvenated *10-year-old* pens!

The best resist pens I've found for circuit-board work are produced in Germany and sold in art stores under the name Staedtler Lumocolor. Medium (no. 317) and fine (no. 318) points are available. (I recommend the no. 318 pen for most circuit-board work.) These pens contain a high-quality waterproof ink and can be opened by removing the top cap (pliers may be necessary in some cases). Most of these pens can be used for circuit board fabrication *without* the solvent-addition treatment just described.

For builders who do not have easy access to an art supply store, I recommend the 0.4-mm, extra-fine-point version of Sanford's® Sharpie® marker. This model has a removable top that allows easy access to the ink cylinder. Many supermarkets stock this pen with stationery supplies or laundry products.

Two types of *medium* point Sharpie pens are available. That labeled PERMANENT MARKER is definitely better for circuit-board work than the no. 3000 "highly water-resistant" model; the permanent marker has the further advantage of easy "openability." (The tip end of the permanent pen is pressed into the barrel assembly portion and held snug with several small rings. If the two parts are simultaneously bent slightly and pulled, the two pieces separate, allowing easy removal of the fiber ink cylinder. Once you've disassembled one of these pens, shave the rings with a file or knife to make subsequent assembly/disassembly easier.) The second-choice (no. 3000) pen is cemented shut; if you must use one of these, I suggest sawing off the top end of the pen to add solvent to the ink cylinder. Reassemble the pen with tape if you do this.

My *ham radio* letter suggests use of a commercial metal-marking lacquer (DYKEM®) as etch resist for the portion of the circuit-board copper intended to remain as a ground plane. If you have trouble locating this product, I recommend thin lacquer, model paint or fingernail polish as a substitute. Be sure the resist you use flows easily so that it can be worked quickly. Also, the resist should be easily removable after etching. (I suggest using acetone as resist-removal solvent.)

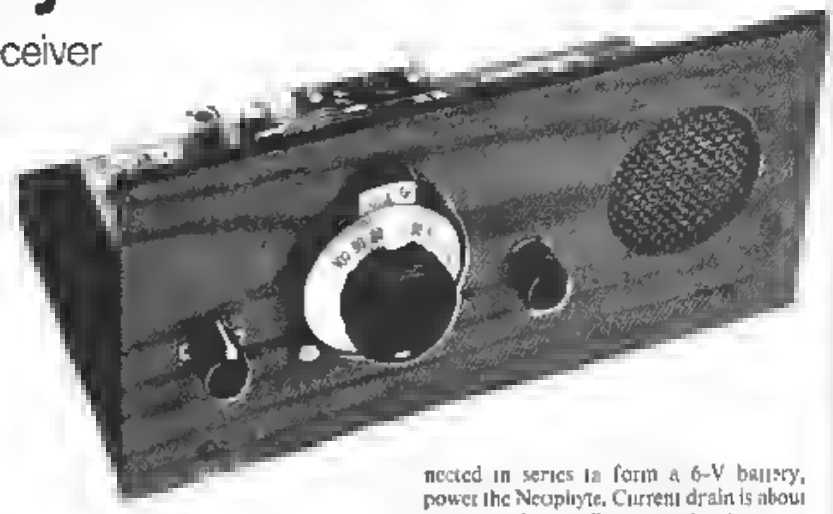
Be sure to take proper safety precautions when working with any of the chemicals I've discussed here: Don't breathe their fumes and keep them out of contact with your skin. Further on the subject of chemicals, I add this. As a retired chemist, I cheerfully object to the characterization of home etched-PC-board fabrication as requiring "messy chemicals." *Chemicals* aren't messy, but *the people who use them* may be!—*Robert J. Grabowski, W5TAP, Rte 1, Box 388, Ozark, AR 72949*

[1]Doug DeMaw, "Homemade Circuit Boards—Don't Fear Them!," *QST*, Aug 1987, pp 14-18.
[2]Robert J. Grabowski, "longhand printed-circuit layout," comments, *ham radio*, Jun 1978, p 6.
[3]Alan Marcus, "A Printed Circuit Board Primer," *CQ*, Oct 1982, pp 44 and 47.

From February 1988 QST, p 14:

The Neophyte Receiver

Looking for a simple receiver to tune the 80- or 40-meter ham bands? Build the Neophyte!



By John Dillon, WA3RNC

Peaktek Electronics
14 Peach Dr
Lewistown, PA 17044

It doesn't take long for prospective hams to discover that there's much more excitement in hearing real signals than listening to "canned" code from a tape or computer program. After all, getting on the air and working with *real* radio is the object of getting an Amateur Radio license! Here is a simple 80- or 40-meter receiver that can bridge the gap between a code-practice machine and your first transceiver by giving you on-the-air listening experience. Dubbed the Neophyte, it's been designed with the needs of the neophyte (beginner) in mind, but will find favor with long-time hams as well.

The Neophyte uses two ICs to receive CW, SSB and AM signals in the 3.5-4.0 or 7.0-7.3 MHz ham bands. It's battery-powered, and most of its circuitry fits on a circuit board just 1-7/8 × 2-5/8 inches in size. The Neophyte's frequency stability allows copy of SSB and CW signals for hours without retuning, and it's sensitive enough to detect signals of less than 0.5 microvolt at its antenna terminals.

How the Receiver Hears

The Neophyte is a *direct-conversion* (D-C) receiver. A D-C receiver converts radio signals directly to audio by mixing the incoming signal with a local oscillator (LO) operating very close in frequency to the incoming signal. The mixing process has this effect: Whenever the LO is tuned so that the frequency difference between it and an incoming radio signal is in the audio range—a few hundred to a few thousand hertz for usable CW, SSB and AM reception—the frequency difference appears at the mixer output as an audio signal. Example: For an incoming Novice CW signal operating at 3737.0 kHz, setting the Neophyte's LO to 3737.6 kHz (a difference of 0.6 kHz, or 600 Hz) will allow you to hear that CW signal as dots and dashes at a 600-Hz pitch. (You could also set the Neophyte's LO to 3736.4 kHz, 600 Hz *below* 3737.0 kHz, to receive the same signal at a 600-Hz pitch.) AM and SSB signals are received by tuning the Neophyte's LO to zero beat—zero frequency difference—with the incoming carrier (or suppressed carrier, in the case of SSB signals). The Neophyte converts the modulation on these signals to audio.

The Neophyte does its D-C job with just two active devices, both of which are ICs. The receiver's *front end*—the RF-handling circuitry from the antenna to the mixer, inclusive—consists of a Signetics NE602N mixer/oscillator IC. The NE602's 8-pin mini-DIP (*mini*ature *d*ual *i*nline *p*ackage) contains bipolar-transistor LO and doubly balanced mixer stages, and a voltage-regulator circuit. The mixer circuitry provides 20 dB of *conversion gain*. This means that the power of an incoming signal is amplified 100 times as the signal is converted to audio by the NE602's mixer and LO.

The other active device in the Neophyte is a National Semiconductor LM386N-1 audio amplifier IC, also contained in an 8-pin mini-DIP. This IC provides 46 dB (power gain, 40,000) of audio amplification to drive headphones or (in a quiet room) a 2¼-inch speaker. Four "C" cells, connected in series to form a 6-V battery, power the Neophyte. Current drain is about 10 mA at low audio-output level.

Fig 1 shows the schematic diagram of the Neophyte. If you'd like to learn the function of each component in the schematic, see the sidebar, "Signal Flow in the Neophyte." You needn't wade through signal flow, however, if you just want to do what we're going to do next: *build* the Neophyte.

Table 1

Neophyte Capacitor Values for 80 and 40 Meters

Band	C1	C7,C8	C9	C10	C11
80 m	330	1000	470	270	120
40 m	not used	330	120	68	150

All capacitances are in pF (1000 pF = 0.001 µF). C1 is disc ceramic; C7-C11 are NP0, polystyrene or silver mica units.

Building the Neophyte

Fig 2 shows a rear view of the Neophyte. Most of the receiver's components are contained on the circuit board. Fig 3 shows the etching pattern for the board, parts placement is shown in Fig 4A. The Neophyte's "cabinet" consists of a 4½- × 5¼-inch piece of ¾-inch-thick pine (base) and a 4- × 8¼-inch piece of ¼-inch-thick particle board, plywood or similar material (front panel). The base can be stained or painted as desired; alternatively, a metal or plastic cabinet can be used to house the Neophyte, if desired.

Components

Although no exotic electronic parts are required, many of the Neophyte's parts are not available at the corner Radio Shack® store. T1 and T2 are 10.7 MHz IF transformers with a 7:1 turns ratio; they have green-colored cores. Other transformers (with different turns ratios) were tried, but receiver performance suffered. Capacitors C7-C11 should be NP0, polystyrene or silver mica units for good frequency stability. At this point, you should decide what band you'd like your Neophyte to cover. The values of C7-C11 depend on the band you choose (see Table 1). For details on the differences between the 80- and 40-meter versions, study the sidebar, "Building the Neophyte for 40 Meters." In the rest of this discussion, I'll concentrate on the construction, testing and adjustment

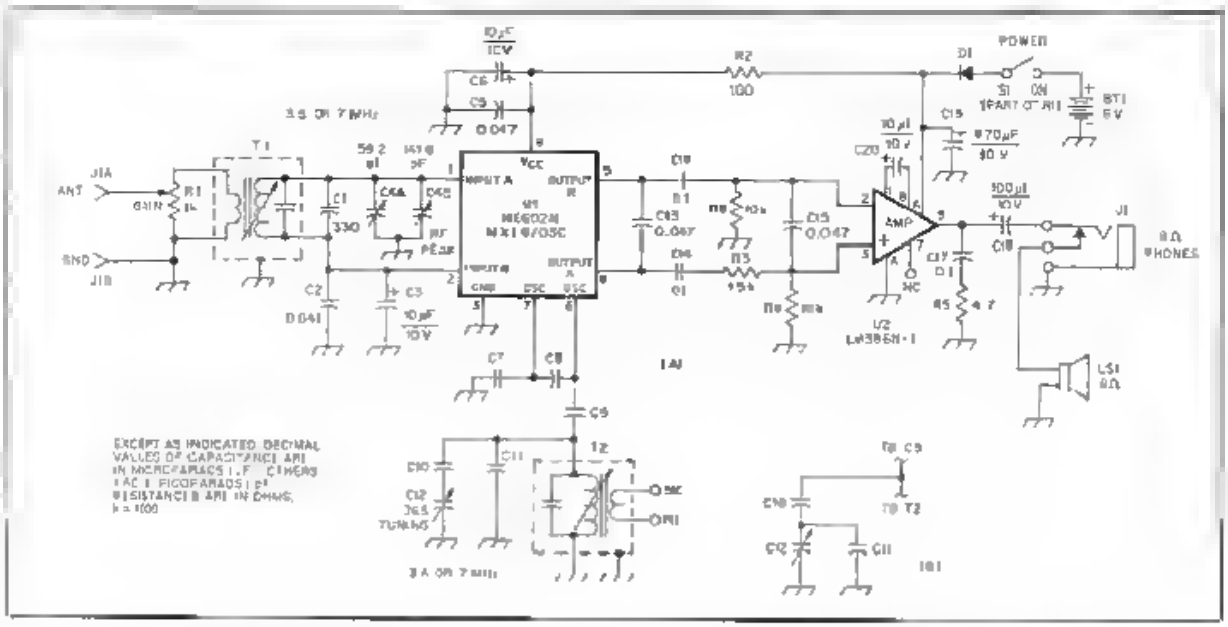


Fig 1—Schematic of the Neophyte receiver. Ceramic capacitors shown below, but not listed in Table 1, may be monolithic or disc units; fixed resistors are ¼-W, carbon film. Component designators shown in the schematic, but not listed below, identify parts for placement on the PC board (see Fig 4). For 40-meter operation, the oscillator circuit is modified slightly, as shown at B. See text and the sidebar, "Building the Neophyte for 40 Meters." Parts kits are available from Pennta< Electronics, see Note 2.

BT1—6 V battery (four "C" cells connected in series).

C1—Ceramic. This capacitor is not used, or is changed in value, for 40-meter operation—see "Building the Neophyte for 40 Meters."

C2, C5, C13, C15—0.047-µF polyester film or ceramic. (0.01 µF also suitable for C2 and C5).

C3, C6, C20—10-µF aluminum electrolytic, 10 to 25 V.

C4—Two-section polyethylene-dielectric variable; sections 59.2 and 141.6 pF (Mouser 24TF222 or equiv). See text.

C7-11—See Table 1.

C12—365-pF, air-dielectric variable (Mouser 524-A1-227, Circuit Specialists A1-227 or equiv). See text.

C14, C16, C17—0.1-µF polyester film or ceramic.

C18—100-µF aluminum electrolytic, 10-25 V.

C19—470-µF aluminum electrolytic, 10-25 V.

D1—1-A, 50-PIV silicon diode (1N4001 suitable).

J1—Two-position terminal strip (Mouser 534-4158, Radio Shack 274-663 or equiv).

J2—Closed-circuit phone jack, 1/8-inch.

LS1—8-Ω speaker, diam 2¼ inches (Mouser 25SPD24 or equiv).

R1—1-kΩ audio-taper potentiometer with SPST switch (Mouser 31VM301 or equiv).

S1—SPST switch mounted on R1.

T1, T2—10.7-MHz IF transformer, 7:1 turns ratio, green core (Mouser 42IF123 or equiv). See text.

U1—Signetics NE602N mixer/oscillator IC (Arrow Electronics 9776CA2).

U2—National Semiconductor LM386N-1 audio-amplifier IC.

Signal Flow in the Neophyte

RF energy from the antenna is fed through GAIN control R1 to the untuned, low-impedance primary winding of T1. This control actually is an RF attenuator. It can be adjusted to prevent very strong signals from overloading U1 when band conditions are especially good, or when strong local signals are present. T1's primary couples incoming signals to the tuned circuit consisting of the T1 secondary, padding capacitor C1, and RF PEAK capacitor C4. This tuned circuit provides *preselection*—it emphasizes signals at its resonant frequency and tends to reject others. The preselected RF is injected into pin 1 of the mixer/oscillator IC, U1.

The oscillator section of U1 serves as the receiver LO. Capacitors C7-C12 and the tuned winding of T2 make up the LO tuned circuit. (T2 is used as a tuned circuit in this application and not as a transformer, its untuned winding is not used.) Because the frequency stability of the oscillator determines the stability of the receiver, temperature-stable capacitors (NP0, polystyrene or silver mica types) are used to minimize drift. Energy is applied to U1 at pin 8. Capacitors C5 and C6 bypass U1's supply pin for ac. Their purpose is to bring the supply pin to ground potential for RF and AF signals while blocking dc. R2 helps these capacitors do their bypassing job by resisting the flow of RF and AF signals on the power-supply line. C2 and C3 are bypass capacitors, also.

Within the mixer section of U1, the LO and preselected RF signals are mixed to provide balanced audio output. The audio appears at pins 4 and 5 of U1. This signal is fed through a simple low-pass filter (C13, C15 and R3) to the inputs (pins 2 and 3) of U2, the audio power amplifier. The low-pass filter tends to pass lower audio frequencies while rejecting higher ones, hence its name. C14 and C16 are blocking capacitors. They block the flow of dc while allowing ac—in this case, audio—to pass.

R4 and R6 set the bias on the input transistors of U2. C20 sets U2's gain to 46 dB. C17 and R5 suppress unwanted HF oscillation in U2. C16 is the output blocking capacitor. Like C14 and C16, it blocks dc while allowing audio signals to pass—in this case, to headphones or speaker. C19 bypasses U2's dc-supply pin for audio. (Because C5, C6, C19 and R2 also work to reduce unwanted audio coupling between U1 and U2 along the dc supply line, they serve as *decoupling* components in the dc line. Decoupling aids stability in high-gain circuits.)

Energy for the Neophyte is provided by four "C" cells connected in series (6 V). S1 is the receiver POWER switch. Diode D1 allows current to pass in only one direction between the battery holder and the receiver circuitry, preventing damage to the receiver components should the batteries be placed in the holder backwards.

About the NE602 Mixer/Oscillator IC

The Signetics NE602 (SA602 for operation over a wider temperature range) is an IC of interest to builders and designers of low-power communications gear, particularly where low power consumption (as during battery operation) is important. Fig A shows its equivalent circuit. The '602 contains doubly balanced mixer, oscillator and voltage regulator elements. Its oscillator circuitry can operate up to 200 MHz in LC and crystal-controlled (fundamental and overtone) configurations. The '602's mixer typically can handle signals up to 500 MHz. Typical dc current drain is 2.4 mA minimum supply voltage is 4.5, maximum 8.0.

The NE602's mixer is known as a *Gilbert cell* multiplier. (If you've ever built a circuit using a Motorola MC1496 or one of its equivalents, you've used a mixer based on the Gilbert cell.) The Gilbert cell consists of balanced switching circuitry driven by a differential amplifier. In the NE602, the amplifier inputs serve as the mixer RF inputs.

The NE602's mixer inputs (RF) and outputs (IF) can be single- or double-ended (balanced) according to design requirements. The resistance of these ports is 1.5 kΩ, the mixer input capacitance is approximately 3 pF up to 50 MHz. The mixer noise figure is typically 5.0 dB at 45 MHz; typical conversion gain is 18 dB at this frequency. The typical two-tone, third-order intercept point of the '602 (measured at 45 MHz with 50-kHz spacing), is −15 dBm.

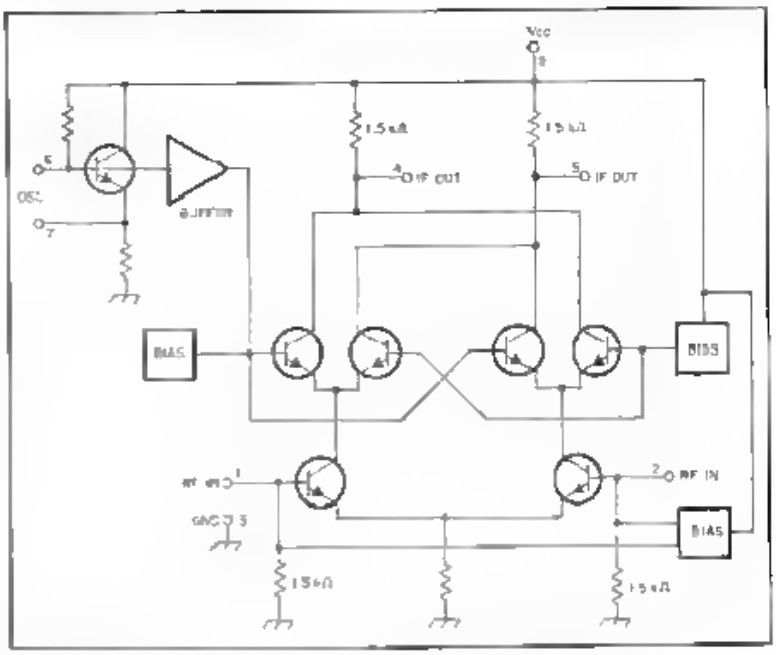


Fig A—The equivalent circuit of the NE602 doubly balanced mixer/oscillator IC.

The Neophyte uses the NE602's on-board oscillator circuitry to achieve good frequency stability at 3.5 and 7 MHz. If the '602's oscillator is unsuitable for a particular application, however, an external LO can be applied to pin 6 of the chip via a dc blocking capacitor. At least 200 mV (P-P) of external-LO drive is required for proper operation of the mixer.—Ed.

This material is based on information in Signetics Corporation's SA/NE602 Product Specification, and in Robert A. Zavrel, "Tomorrow's Receivers: What Will the Next Twenty Years Bring?," *Ham Radio*, Nov '987, pp 8-9, 11-13 and 15.

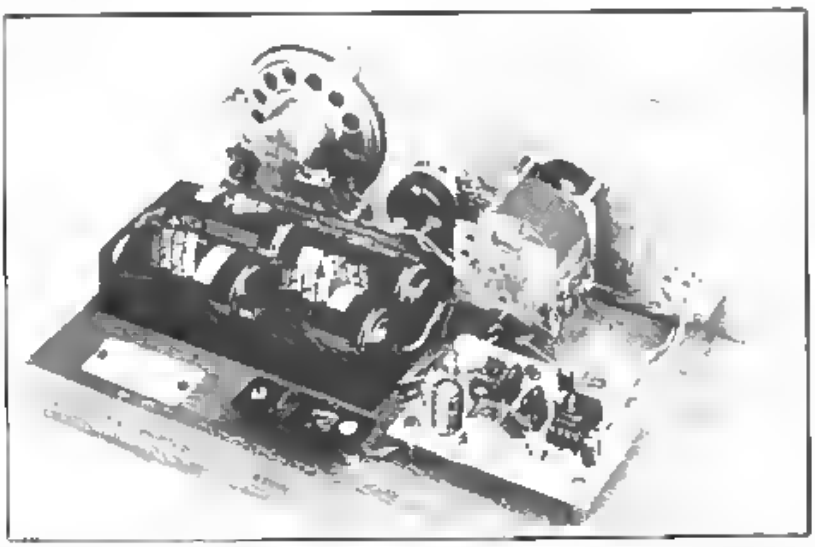


Fig 2—The Neophyte's cabinet, battery and front-panel controls dwarf its circuit board (right foreground). From left to right, the front-panel components are J2, LS1, R1, C12 and C4. The tuning capacitor mounts to the front panel by means of flat-head, 1-inch, no. 6-32 screws and no. 6-32 nuts. The screws serve both as fasteners and mounting standoffs for the tuning capacitor (see Fig 5). The antenna terminals have been colored with felt-tip markers to indicate their function: black for the ground connection, green for the antenna.

of the 80-meter Neophyte.

The TUNING capacitor, C12, is a 365-pF, air-dielectric unit. One section of a two-section capacitor can be used at C12, but you may have some difficulty mounting such a capacitor to the front panel with the technique we'll cover shortly in "Construction." Local hamfests offer an excellent opportunity to find a tuning capacitor, as well as a reduction drive to turn it. These reduction drives are generally found in two diameters, 1½ and 2 inches. You can use either size.[1]

RF PEAK capacitor C4 is a two-section, plastic-dielectric variable. Similar capacitors are commonly used in small, portable radios. The unit specified for C4 in the parts list (see the Fig 1 caption) provides a maximum capacitance of about 200 pF with both sections connected in parallel. Off the shaft, its shaft is only about ¼ inch long—too short to be useful for our purposes. The shaft can be extended by bolting a ¼-inch-diameter round standoff, ½ inch long, to the existing C4 shaft. Use a *metric* (2.5-mm-diameter) screw to match the threads in C4's shaft. The force necessary to turn a non-

[1]Footnotes appear at end of article.

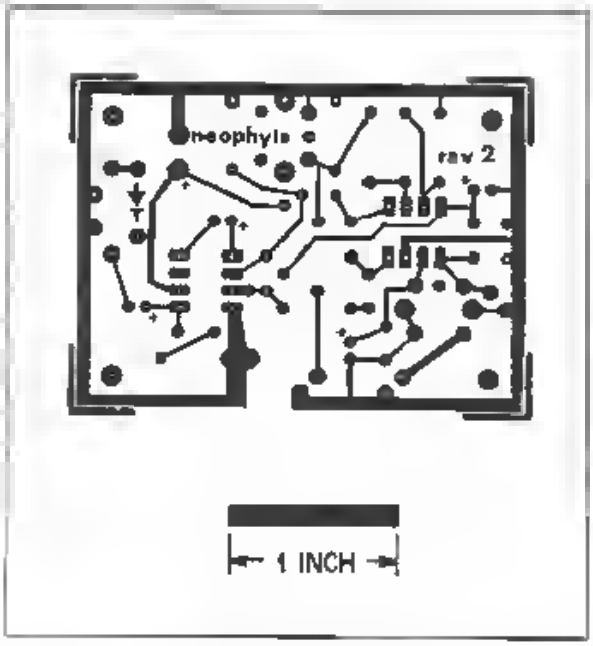


Fig 3—Circuit-board etching pattern for the Neophyte. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil.

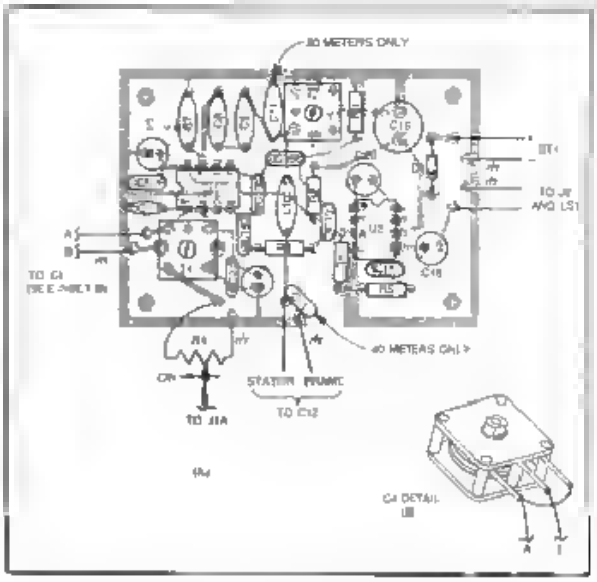


Fig 4—Parts-placement guide for the Neophyte (A) and detail of connections to C4 (B). Parts are placed on the nonfoil side of the board. The shaded area represents an X-ray view of the copper pattern. The placement of C11 depends on whether 80- or 40-meter coverage is desired; see text and "Building the Neophyte for 40 Meters."

metric screw into C4 can destroy the capacitor. (By the way, don't be tempted to use one of these inexpensive plastic capacitors for C12, the TUNING capacitor. You would be disappointed with the tuning drift that occurs as the capacitor's dielectric sheets settle each time you tune the receiver.)

In general, it's best not to attempt parts substitutions. By using the specified parts, you stand the best chance of being rewarded with a receiver that works correctly the first time it's turned on. Etched and drilled PC boards, and complete parts kits, are available from Penntek Electronics.[2] The Appendix shows the addresses of parts distributors if you'd rather order direct from them. Note, however, that some of these firms may have minimum order requirements or small-order service charges.

Construction

Building the cabinet and mounting controls and mechanical components is the greater part of constructing the Neophyte, so do this job first! Mounting the TUNING capacitor and reduction drive to the panel is the most time-consuming part of construction. The panel hole for the reduction-drive bushing must be large enough to allow rotation of the bushing and its set screw, but small enough to leave enough material to pass and hold the flat head screws used to mount the TUNING capacitor (see Fig 5). The best way to mark these holes is to make a drilling template by pushing a piece of paper down over the capacitor shaft. The shaft punches through the paper, marking the position of the hole for the reduction-drive bushing. Next, hold the paper against the capacitor frame and use a pencil point to punch holes in the paper corresponding to the mounting-screw holes in the capacitor frame. Instant drilling template! The bushing hole shown in Fig 5, 7/8 inch in diameter, leaves just enough panel material to hold the countersunk holes for the three no. 6-32 capacitor mounting screws. The best technique is to enlarge the bushing hole *last*, widening it only enough to pass the

Building the Neophyte for 40 Meters

The Neophyte receiver can be built for 7.0-7.3 MHz coverage as follows: Omit C1, C7 through C11 take the 40-meter values shown in Table 1. C11 is mounted in parallel with C12 instead of across T2 (see Fig 1B); this is easily done by mounting C11 across the PC-board connections to C12. Before mounting T2 to the circuit board, remove the small, tubular capacitor in the base of the transformer. Do this carefully with a small razor knife.

Forty-meter alignment is similar to that for the 80-meter Neophyte. Adjust T2 for an oscillator tuning range of 7.0-7.3 MHz, with some overtravel at both ends of the range. With the TUNING control set to the center of the band, set the RF PEAK knob to one o'clock. Adjust T1 for maximum signal strength. This completes alignment of the 40-meter Neophyte.

Because of decreased LO-mixer isolation in the NE602 at 7 MHz, adjustment of the RF PEAK control "pulls" the LO slightly in the 40-meter Neophyte. (Pulling is perceptible as a shift of received-signal pitch as RF PEAK is varied.) This isn't much of a problem, because the RF PEAK control needs little adjustment from one end of the 40-meter band to the other. In fact, you can eliminate the RF PEAK control in the 40-meter Neophyte if you do most of your listening in one part of the band. To do this, omit C4, install a 150-pF capacitor at C1 and adjust T1 for maximum signal strength at your favorite spot in the band.

The Neophyte in ARRL Lab Tests

ARRL Lab testing of one sample of the 80-meter Neophyte netted these results: minimum discernible signal (MDS), −118 dBm (decibels relative to a milliwatt) at 3520 kHz and −113 dBm at 3747 kHz; two-tone, third-order dynamic range with 100-kHz tone spacing 73.5 dB; selectivity, 1 kHz at −3 dB and 7.5 kHz at −20 dB. Blocking dynamic range was not measured. No microphonics were noted.

The frequency coverage of the sample receiver was 3473-4027 kHz. The poorer of the two MDS figures above (−113 dBm) confirms that the Neophyte is capable of detecting signals down to 0.5 microvolts across 50 ohms, as specified by WA3RNC. At 3520 kHz, sensitivity improved to just under 0.3 microvolt.—Ed.

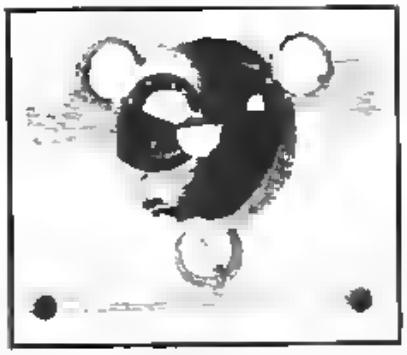


Fig 5—Detail of the mounting holes for the TUNING capacitor, C12. The holes are countersunk to keep the screw heads flush with the panel surface, allowing the reduction drive to be mounted flat to the panel. See text.

reduction-drive bushing and its set screw—*after* the capacitor mounting screw holes have been drilled.

Mounting the Neophyte's TUNING capacitor to the front panel as just described provides good mechanical stability. If you use a capacitor other than the one shown in the parts list for C12, carefully study your mounting options. Make sure that the mounting screws don't penetrate too far into the capacitor and damage the plates. Whatever you use for C12, mount it to the panel rather than the base, if possible.

The specified speaker requires a 2-inch hole (or a field of ¼-inch holes 2 inches in diameter). A small piece of wire screen or grille cloth between panel and speaker can be used to protect the speaker cone. (This is especially important if you mount the speaker behind a single large hole.) The speaker is attached to the panel with clips designed for this purpose (Mouser Electronics 48SC004); hot-melt glue or epoxy cement can be used instead. A 3.5-mm or 1/8-inch headphone jack is used to allow connection of inexpensive transistor radio earphones; if you prefer, you can use a ¼-inch jack here.

The antenna connectors (J1A and J1B in Fig 1) are part of a two-position terminal strip, this is mounted to the receiver base by means of standoffs and screws. Radio Shack push-button speaker terminals would be a good substitute here. The battery holder is a 4-"C"-cell holder from Mouser or Radio Shack.

Solder the components to the circuit board, being careful to observe capacitor polarity and IC orientation. (I recommend that you use IC sockets instead of soldering the ICs directly to the board.) After you've soldered the components to the board, cut off excess wire. Check carefully for solder bridges between circuit-board traces, proper electrolytic capacitor polarity, and correct orientation of D1, U1 and U2. If all looks well, wire the board into the rest of the receiver. As shown in Fig 2, use twisted-pair wiring for connections to C4, J1, J2 and R1. The capacitor specified for C4 has three terminals. Fig 4B shows how to wire these for connection to the circuit board.

When you've completed all connections, mount the board to the cabinet base by means of screws and spacers. Next, we'll align and test the Neophyte.

Checkout and Alignment

Before applying power to the receiver, recheck your wiring once again. Install four "C" cells in the Neophyte's battery holder. (Note: You can use a regulated dc supply in place of the batteries if you wish, but *do not apply more than 9 V to the receiver*, or you'll damage the ICs.) Install a milliammeter or digital multimeter (DMM) in series with the batteries, and turn on the receiver. If the meter indicates less than 15 mA, all's well so far.

Adjust the TUNING capacitor almost to minimum capacitance (plates just short of fully unmeshed). Connect a signal generator to the antenna terminals and inject a 500-µV, 4-MHz signal into the Neophyte.[1] Turn the Neophyte's GAIN control to maximum (fully clockwise if you've wired it correctly) and adjust oscillator coil T2 until you hear the test signal.

Position the RF PEAK knob on C4's shaft so that maximum capacitance (knob fully counterclockwise) is at nine o'clock and minimum capacitance is at three o'clock. Set the RF PEAK capacitor nearly to minimum capacitance (almost fully clockwise, near two o'clock) and adjust T1 for maximum signal strength. Verify that the receiver tunes 3.5-4.0 MHz with a slight overtravel at both ends of the range. Also check that the RF PEAK control tunes through resonance at both ends of the band.

Disconnect the signal generator from the Neophyte and connect a good antenna, such as a dipole, to the receiver. As you tune the Neophyte across the band, adjust the RF PEAK control for best signal strength. (Don't expect outstanding performance with a clip-lead antenna!) If you don't have a dipole, use a long random-wire antenna. (Use of a random-wire antenna also requires a ground connection.) Set the GAIN control no higher than necessary for solid reception, this reduces the likelihood of detector overload. This practice also lengthens battery life because U2 draws more energy from the battery as the receiver output increases. Battery life, longest when headphones are used in place of the speaker, can exceed 300 hours when fresh alkaline cells are used!

Summary

I welcome your comments and questions on the Neophyte—please include an SASE if you expect a reply. Several Neophytes have been built using different construction techniques. All perform flawlessly. The Neophyte usually can hear any signal audible on a typical ham transceiver. Its selectivity is adequate for band scans and casual listening, and it's an excellent project for schools, ham-radio classes, beginners and old-timers. In short, the Neophyte is fun!

APPENDIX

Parts for the Neophyte are available from a combination of these sources, and from Pennlek Electronics (see Note 2).

Arrow Electronics
25 Hub Dr
Melville, NY 11747
tel 800-932-7769

Mouser Electronics
2401 Hwy 287 N
Mansfield, TX 76063
tel 817-483-4422

Circuit Specialists
PO Box 3047
Scottsdale, AZ 85257
tel 602-966-0764

Radiokit
PO Box 973
Pelham, NH 03048
tel 603-635-2235

Mouser Electronics
11433 Woodside Ave
Santee, CA 92071
tel 619-448-2222

Notes

[1]Mouser Electronics carries 1⅛- and 2-inch reduction drives as part nos. 45KN100 and 556-S50, respectively. Radiokit carries a 2-inch drive as part no. S-50. See the Appendix for the addresses of these firms.

[2]Circuit boards and parts kits for the Neophyte receiver are available from Pennlek Electronics, 14 Peace Dr, Lewistown, PA 17044, tel 717-248-2507. Prices are as follows: (1) An etched and drilled PC board, $4.50; (2) all PC-board-mounted components, and an etched and drilled PC board, $17.50; (3) a complete Neophyte kit, including drilled wooden panel, wooden base, all hardware and parts, $45. (4) builders who wish to supply their own parts for the Neophyte, but who have trouble finding a small-quantity source for the Signetics NE602N IC, can purchase the NE602N from Pennlek Electronics for $3.25 postpaid. Add $3.50 for shipping and handling to all orders for options 1, 2 and 3. Pennsylvania residents, add sales tax to all orders. When ordering options 2 and 3, be sure to specify 80- or 40-meter operation. The ARRL and QST in no way warrant this offer.

[3]If you don't have access to a signal generator, you may be able to generate a test signal by feeding a 4-MHz transmitter signal into a dummy load. Connect a short length of wire to the Neophyte's antenna terminal (J1A) and bring the wire near the dummy load. Vary the spacing between the wire and the load—or reduce the transmitter output—until the transmitter signal is just strong enough to use.—Ed

For updated supplier addresses, see ARRL Parts Supplier List in Chapter 2.

Microphonics, Hum, LO Radiation: Low to Absent in the Neophyte

Although the Neophyte is simple, it does not exhibit the drawbacks sometimes associated with D-C receivers. Microphonics—unwanted noises that occur with vibration when electronic parts unexpectedly act like microphones—simply do not exist with this receiver, and I have not heard any hum, even when using an external power supply. LO radiation—sometimes a problem in simple receivers using LOs at any frequency—*isn't* a problem with the Neophyte. The measured LO level at the receiver antenna terminals is only 80 microvolts.

A Band-Imaging CW Receiver for 10 and 18 MHz

Fig 1 — The band-imaging receiver covers the CW segments of the 10- and 18-MHz amateur bands with good stability, sensitivity and single-signal selectivity. Larger controls are (L) TUNING, IF GAIN and AF GAIN. The two smaller controls, separated by the PHONES jack are BAND and SIDETONE LEVEL. The tuning scale is drawn on contact paper applied to an aluminum disc.

Band Imaging has long been used in Amateur Radio as a means of making a stable local oscillator (LO) do double duty. Instead of building equipment using only one LO-to-RF relationship for frequency conversion—with, for instance, only the *difference* between the LO and higher-frequency incoming signals giving output at the IF—*two* of several LO-to-RF relationships can be exploited for two-band coverage. A band-imaging receiver appeared in every edition of this *Handbook* from 1953 through 1966, from "A Two-Band Four-Tube Superheterodyne" in 1953 to "The HB-65 Five-Band Receiver" in 1966. Each of these receivers converted the 80- and 40-meter amateur bands to a 3.5-MHz IF by means of a 5.2-5.7-MHz LO. On 80 meters, the conversion relationship in each receiver is LO − RF = IF; on 40, the relationship is RF − LO = IF. Both bands "tune in the same direction" with this system: The received frequencies of 3.5 and 6.9 MHz correspond to the lower limit of the LO tuning range.

Band Imaging can also be used to cover the 80- and 20-meter amateur bands: A 5.0-5.5-MHz LO is used to convert each band to a 9-MHz IF. In each system, the LO-to-RF relationship on 80 meters is RF + LO = IF, on 20, RF − LO = IF. The drawback to this band-imaging system is that the lower band "tunes backwards". The lower limit of the LO tuning range corresponds to 4.0 MHz on 80 meters and 14.0 MHz on 20. Nonetheless, the 80/20 band-imaging system has also been popular with radio amateurs because of the inherent sideband inversion between the image bands: The BFO-to-IF relationship that affords LSB reception on 80 meters demodulates USB on 20.

With this overview of band-imaging technique in place, we present a band-imaging CW receiver for 10 and 18 MHz (see Fig. 1). Using a 14-MHz LO, it converts the entire 10-MHz amateur band, and the CW portion of the 18-MHz amateur band, to a 4-MHz IF. Both bands tune in the same direction. At 4 MHz, a four-crystal ladder filter provides single-signal selectivity. The design emphasizes good basic receiver performance with an eye toward compactness; hence, features such as a digital frequency display, AGC and active audio filtering have been omitted. Alignment and checkout of the band-imaging receiver requires only (1) a 51-ohm resistor; (2) a receiver capable of CW reception at 14.0-14.2 MHz and 4 MHz ±1 kHz with an S meter and frequency display resolution of 1 kHz or greater; and (3) a crystal-controlled marker generator capable of providing 10-kHz markers. The performance measurements given later in this article were obtained from a receiver aligned *by ear* with such test equipment. You need not have access to a radio lab to enjoy similar results. David Newkirk, AK7M, designed and built the project in the ARRL lab.

Circuit Description: RF Amplifiers

A separate 40673 RF amplifier is used for each band. (See Fig. 2.) The circuit is electrically identical to that used for the RF amplifiers in "A High-Performance Communications Receiver," presented later in this chapter. Several other circuits in the band-imaging receiver are based on the K5IRK/W7ZOI high performance design. To simplify alignment of the band-imaging

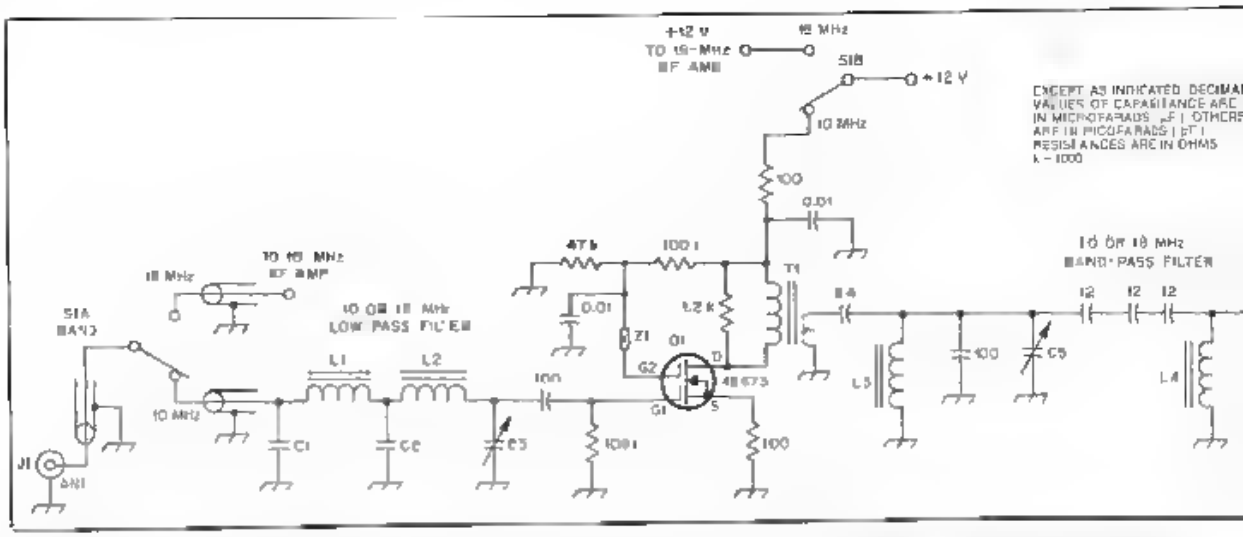


Fig 2 — Schematic of the RF amplifiers for the band-imaging receiver. A separate amplifier is used for each band. The low- and band-pass filters may be aligned with the aid of a crystal-controlled marker generator; see text. Capacitors are disc ceramic unless otherwise noted. Capacitors marked with polarity are electrolytic. All resistors are ¼-W, 10% units unless otherwise noted.

C1, C2, C4, C7 — Silver mica, polystyrene or ceramic capacitor, see Table 1 for values.
C3, C5, C6 — Ceramic or mica compression trimmer. Mouser Electronics ceramic trimmer 24AA067 (12-100 pF) used for 100 pF. Mouser 24AA064 (5-45 pF) used for 45 pF.
J1 — Coaxial RF connector.
L1-L4 — Wound on Amidon T-50-6 powdered-iron toroid core or equiv. All inductors use no. 22 enameled wire with one exception. For L2 at 10 MHz, use no. 24 enameled wire. See Table 1 for number of turns.
Q1 — 40673 dual-gate MOSFET.
S — 3PDT toggle (Radio Shack 275-661 or equiv).
T1 — Transformer wound with no. 28 enameled wire on Amidon FT-37-43 ferrite toroid core or equiv. Primary (168 µH), 20 turns; secondary (6.7 µH), 4 turns.
Z1 — Ferrite bead on Gate 2 lead of Q1, Amidon FB-43-101 or equiv.

Table 1

Component Values for the Band-Imaging Receiver RF Amplifiers

MHz	C1	L1	C2	L2	C3	C4, C7	L3, L4	C5, C6
10	300	13	680	28	100	33	17	100
18	180	10	390	22	45	22	10	100

Values listed for capacitors are capacitance in pF. Values listed for inductors are number of turns of wire required.

receiver, the variable coupling capacitor (C15 in Ch. 30, Fig. 12) between the two sections of the output filter is replaced by three 12 pF capacitors in series. Gain of this circuit is 12 to 15 dB, depending on alignment and the characteristics of Q1. Band changing is accomplished by switching RF input, RF output and dc connections between the 10- and 18-MHz amplifier boards via S1, a 3PDT toggle. Input and output (I/O) impedances of each RF amplifier board are 50 ohms.

Mixer, IF Filter and IF Amplifiers

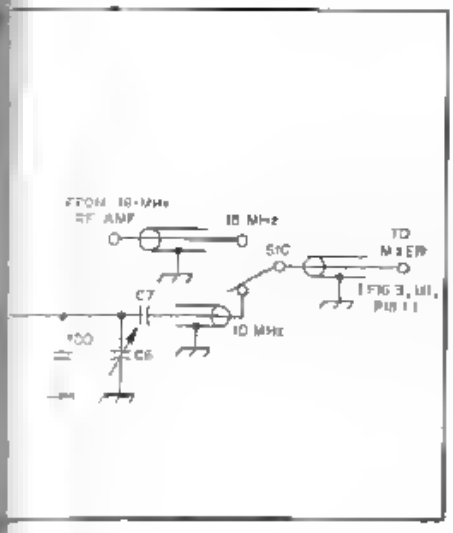
See Fig. 3. The band-imaging receiver uses a Mini-Circuits SBL-1 doubly balanced diode-ring mixer (U1) followed by a strong bipolar-transistor IF amplifier (Q2).[1] This is the circuit used in the K5IRK/W7ZOI receiver, with several modifications. In the band-imaging receiver, the bifilar 4:1 collector transformer in the original design has been replaced with a toroidal monofilar choke, RFC1. The supply end of the 1-kΩ Q2 base bias resistor is now connected directly to the 12-V dc line at the cold end of RFC1. This removes the RF feedback present in the original circuit. Surprisingly, this feedbackless configuration results in better sensitivity *and* two- one 3rd-order IMD dynamic range than the unmodified circuit affords *at a 4-MHz IF*. The original circuit, intended for use at an IF of 9 MHz, did not provide a comparable performance even when the inductance of its 4:1 collector transformer was scaled for 4 MHz.

The post-mixer amplifier feeds a four-crystal ladder filter via a 6-dB pad. The I/O impedances of the crystal filter are 200 ohms. Because this is a good match for the collector impedance of Q2, the step-down transformer in the original post-mixer amplifier circuit is not required. The 50-ohm 6-dB attenuator of the original circuit has been scaled to 200 ohms. This pad should *not* be replaced with one of lower attenuation: It assures a nonreactive wideband termination for Q2 *and* the crystal filter. Less attenuation here results in reduced IMD dynamic range, as confirmed by lab tests.

[1]Mini-Circuits, PO Box 166, Brooklyn, NY 11235, tel 212-934-4500.

The crystal filter was designed using Hayward's technique (see "Simple Cohn Crystal Filters," *QST*, July 1987, pp 24-29). Measured selectivity of the prototype filter was 405 Hz at −6 dB and 1650 Hz at −60 dB, resulting in a −60 dB/−6 dB shape factor of 4.57. Insertion loss was 2 dB, and passband ripple was less than 0.4 dB. As is characteristic of simple ladder crystal filters, the upper passband slope is steeper. Because of this, the BFO must be set on the *upper* side of the filter for best single-signal reception. With the BFO set to provide a 550 Hz beat note for signals at IF center, rejection of the audio image in the prototype receiver was 73 dB. Ultimate attenuation was 90 dB.

No filter adjustment is necessary, but it *is* important that you use the specified crystals if you intend to duplicate the post-mixer-amplifier/pad/filter arrangement shown in Fig. 3. Substitutions at Y1-Y4 will require filter capacitors of other than 300 pF, resulting in I/O impedances of other than 200 ohms. Hayward states that the series-resonant frequencies of the four filter crystals must fall within a spread of no more than 10% of the desired *3-dB* filter bandwidth. We chose to evaluate the performance of the filter in the more popular terms of −60 and −6 dB bandwidths, it follows that 10% spread is too generous when a given −6 *dB* filter bandwidth is the target. Experiments with various new and surplus 4-MHz microprocessor-clock crystals in the ARRL lab showed that the new International Crystal Mfg. (ICM) crystals provided the best



performance overall. Shape factors (– 60 dB/– 6 dB) for the clock-crystal filters were rarely less than 5, and sometimes more than 6. I/O impedances were between 300 and 400 ohms. Several times, four crystals within a suitably narrow frequency spread could be found only by grading *10 or more* clock crystals. Custom-ground crystals offer the added advantage of resonating —within tolerance, of course—on the frequency you specify. Their unit price is higher, but they come closest to guaranteeing that your filter will perform as predicted.

Post-filter IF amplification is provided by U2, an MC1350P video amplifier IC. The 200-ohm resistor between pins 4 and 6 of U2, in conjunction with the 0.1 µF bypass capacitor at pin 6, terminates the crystal filter output. Manual gain control is achieved by applying a variable positive voltage to pin 5 of U2 through a 27-kΩ resistor and IF GAIN control R1. Receiver muting is accomplished by means of Q3. Grounding the MUTE terminal (center conductor at J2) applies maximum gain-reduction voltage to U2. The supply voltage (nominally 12) appears across J2 with the receiver unmuted; current through the grounded MUTE line is 5 mA. IF output (Z = 50 ohms) is available at the secondary of T2.

Local Oscillator

The schematic of the band-imaging receiver LO is shown in Fig. 4. An MPF102 JFET, Q4, operates as a Colpitts oscillator. The oscillator signal is amplified by Q5, a 40673 dual-gate MOSFET. Bandspread is achieved by tapping the tuning capacitor, C9, down on LO tank inductor L5. Tuning range of the circuit is approximately 14.060 to 14.153 MHz. An air-dielectric trimmer, C10, shifts this range for dial calibration.

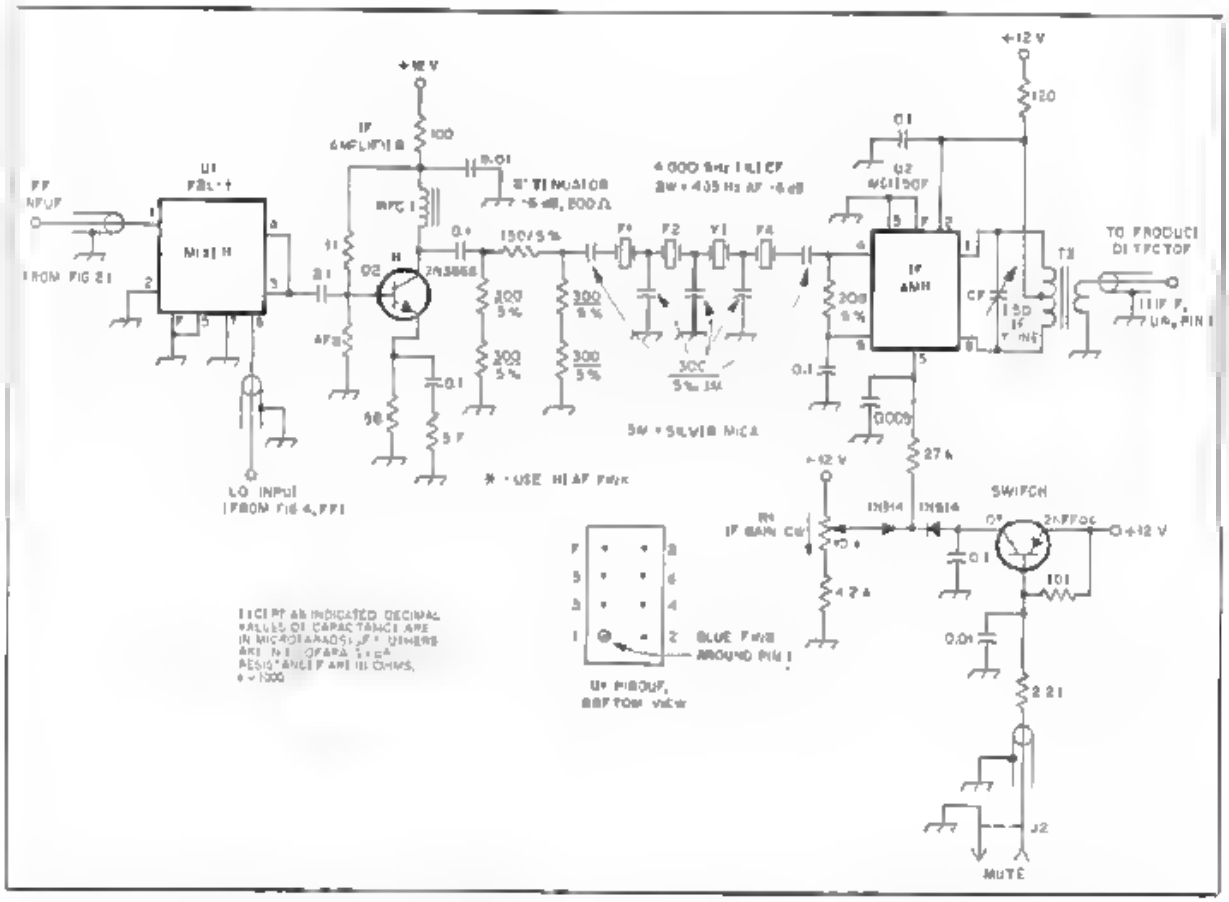


Fig. 3 — Schematic of the mixer, crystal filter and IF amplifier stages of the band-imaging receiver. Capacitors are disc ceramic unless otherwise noted. Capacitors marked with polarity are electrolytic. All resistors are ¼-W, 10% units unless otherwise noted.

C8 — 360-pF compression trimmer (Arco 428 or equiv).
J2 — Phono jack.
Q2 — 2N3866 or 2N5109. Use a small heat sink on this transistor.
Q3 — 2N3906.
R1 — 10-kΩ linear potentiometer.
RFC1 — 95 µH, 15 turns no. 24 enameled wire on Amidon FT-37-43 ferrite toroid core.
T2 — Transformer wound on Amidon T-68-1 powered-iron toroid core, or equiv. Primary (12.9 µH), 36 turns no. 26 enameled wire, center tapped; secondary (0.9 µH), 3 turns no. 26 enameled wire over center of primary.
U1 — Mini-Circuits SBL-1 doubly balanced diode-ring mixer.
U2 — MC1350P video amplifier IC.
Y1-Y4 — 4.000000-MHz custom-etched crystal, 26°C calibration temperature, grade CS-1 (0.001% tolerance), HC-6/U holder, series resonant, International Crystal Mfg. Co. type 433340. See text.

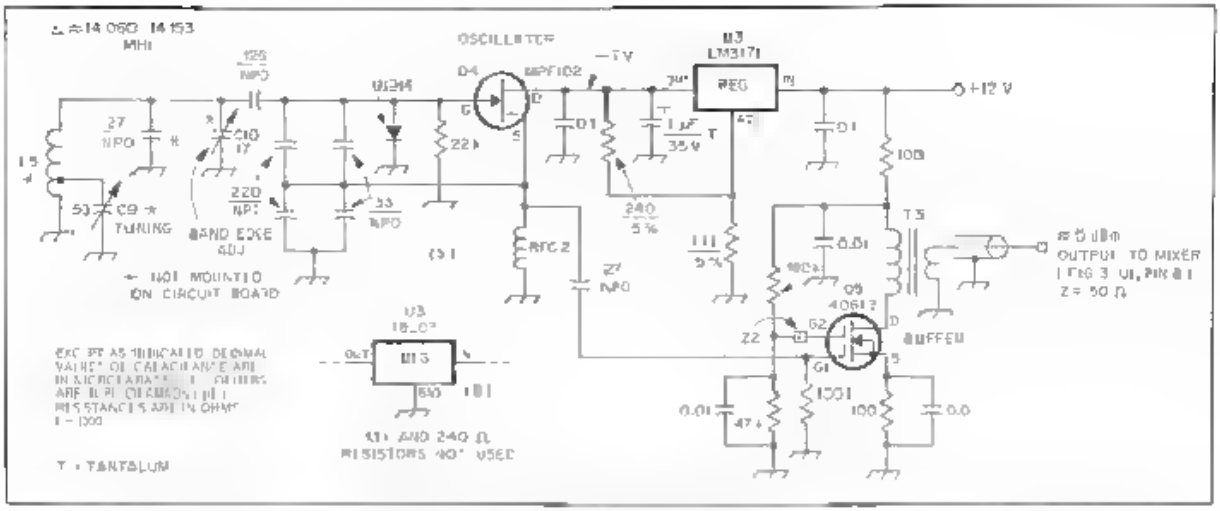


Fig 4 — Schematic of the band-imaging receiver LO and buffer circuit. Capacitors are disc ceramic unless otherwise noted. Capacitors marked with polarity are electrolytic. All resistors are ¼-W, 10% units unless otherwise noted. At A, an LM317L adjustable regulator is used at U3. The inset at B shows connections for an 78L07 regulator at U3. For best stability, use only NP0 (C0G) capacitors in the circuitry associated with the gate and source leads of Q4. Space L5 by at least its diameter from other components and the LO shield box. See text and Fig 8.

C9 — 50-pF air variable (Jackson Bros 4857-50 or equiv)
C10 — 17-pF air trimmer (Johnson 189-506-5 or equiv)
L5 — 1.4 µH, 11 turns no. 22 tinned wire, 24 turns per inch (Barker & Williamson 3036 Miniductor). Tap at 2 or 3 turns from ground end. See text and Figs 8 and 8B.
Q4 — MPF102 JFET
Q5 — 40673 dual-gate MOSFET
RFC2 — 39 µH Miller 70F395AS, or 24 turns no. 26 enameled wire on Amidon FT-50-61 ferrite toroid core.
T5 — Transformer wound on Amidon FT-37-43 ferrite toroid core or equiv. Primary 150 µH, 11 turns no. 26 enameled wire. Secondary 3.8 µH, 3 turns no. 26 enameled wire.
U3 — Voltage regulator, LM317LH, LM317LZ or (with circuit changes shown at inset B) 78L07.
Z2 — Ferrite bead on Gate 2 lead of Q2. Amidon FB-43-101 or equiv.

Despite the relatively high LO operating frequency, stability is good. Measured drift of the point-to-point-wired prototype oscillator was 530 Hz in the 45 minute period after turn-on, 460 Hz of which occurred in the first ten minutes. Over the next three hours, this oscillator drifted approximately *20 Hz*. Stability was even better with the circuit rebuilt on an etched circuit board: Drift for the ten minutes after turn-on was only 256 Hz. *The key to this stability is the use of NP0 (C0G) ceramic units for all fixed capacitors associated with the gate and source of Q4.* Although silver-mica or polystyrene capacitors may be hand-picked for low drift, only NP0 capacitors offer minimum drift "off the shelf." Oscillator stability is further improved by the use of a three-terminal regulator to stabilize Q4's drain supply, and by enclosing LO and buffer in a shield box to slow the effect of changes in ambient air temperature.

BFO, Detector and Audio Stages

The K5IRK/W7ZOI crystal-controlled BFO is used in this receiver with one modification. The secondary of T4 in Fig. 5 carries only RF and no dc. Y5 is an inexpensive 4-MHz microprocessor clock crystal. Every such crystal we tried worked well in this circuit; a custom-ground crystal is unnecessary here.

The detector and AF stages of the band-imaging receiver are shown in Fig. 6. The product detector (U4) is a Mini-Circuits SBL-1 doubly balanced diode-ring mixer. RFC3 and the 0.001-µF capacitor provide RF filtering ahead of the AF preamp, U5, an NE5534 low-noise audio op amp. The parts list for Fig. 6 specifies a "wind-it-yourself" toroidal choke for RFC3; pi-wound chokes tried here were prone to pickup of 60-Hz harmonics.

U6, an LM380N-8, serves as the AF power amplifier. Its output is connected to a front-panel stereo headphone jack, J4, and a rear-panel phono connector, J5. J4 is wired to accept stereo headphones, monaural phones may be used if inserted no farther than the first detent. The 1-kΩ resistor from the output lead to ground serves to charge U6's 470-µF output coupling capacitor at power-up if a head-

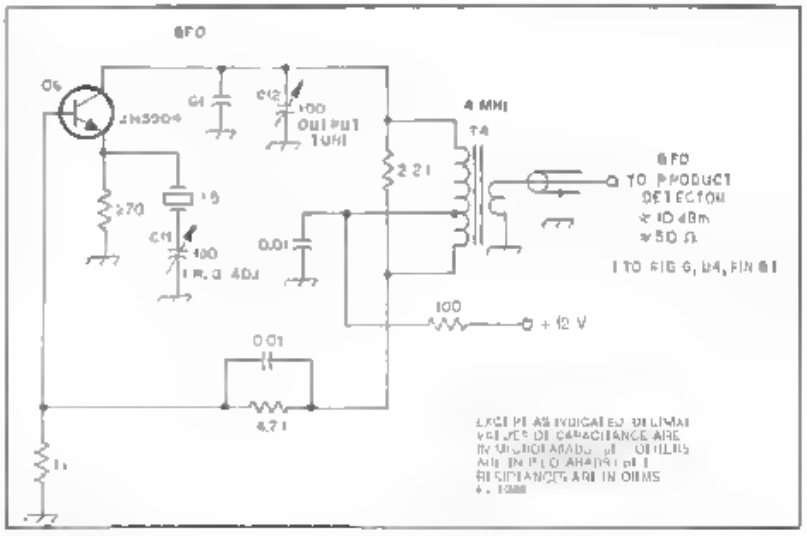


Fig 5 — Schematic of the band-imaging receiver BFO. Capacitors are disc ceramic unless otherwise noted. Capacitors marked with polarity are electrolytic. All resistors are ¼-W, 10% units unless otherwise noted.

C11, C12 — 100-pF ceramic or mica compression trimmer. Mouser Electronics 24AA067 (12-100 pF) suitable.
Q6 — 2N3904
T4 — Transformer wound on Amidon T-68-2 powdered-iron toroid core or equiv. Primary (19.8 µH): 59 turns no. 28 enameled wire, tapped at 12 turns. Secondary (0.98 µH): 8 turns no. 28 enameled wire over cold end of primary.

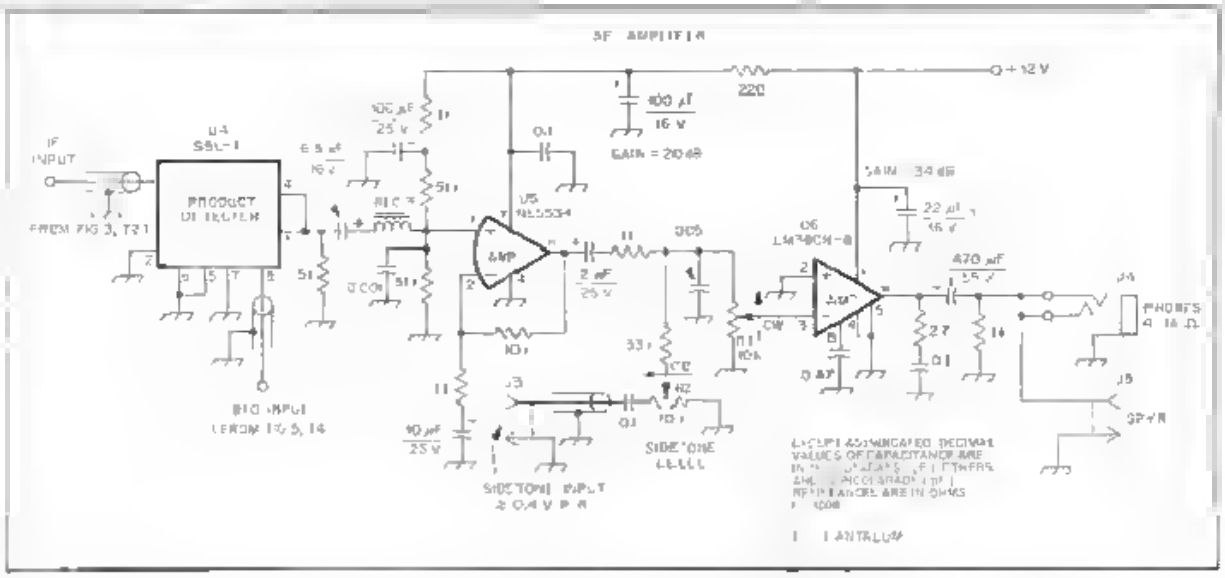


Fig. 6 — Schematic of the product detector and audio amplifiers for the band-imaging receiver. Capacitors are disc ceramic unless otherwise noted. Capacitors marked with polarity are electrolytic. All resistors are ¼-W, 10% units unless otherwise noted.

J3, J5 — Phono jack.
J4 — Stereo headphone jack.
R2, R3 — 10 kΩ audio taper potentiometer.
RFC3 — 1-mH RF choke. 34 turns no. 30 enameled wire on Amidon FT-37-72 ferrite toroid core or equiv.
U4 — Mini-Circuits SBL-1 doubly balanced diode ring mixer.
U5 — NE5534 low-noise audio op amp.
U6 — LM380N-8 audio power amp.

phone or speaker load has not already been installed at J4 or J5. Without this resistor, the capacitor would charge on connection of the audio transducer, resulting in a loud "thump."

As mentioned earlier, no active audio filtering is included in the band-imaging receiver. The higher audio components in detected IF amplifier hiss are reduced by the 0.05-µF capacitor connected between the hot end of the AF GAIN control, R3, and ground.

Sidetone can be injected into the audio chain at J3. Sidetone level is adjusted from the front panel by R2. Setting R2 to minimum shunts the AF GAIN control with a 33 kΩ resistor. This reduces overall audio gain by less than 1 dB. A 400-mV signal at J3 provides more than enough sidetone audio at normal AF GAIN settings.

This receiver requires, but does not include, a regulated dc power supply capable of providing a maximum of 220 mA at 12 V. See Chapter 27, Power Supply Projects, for suitable circuits.

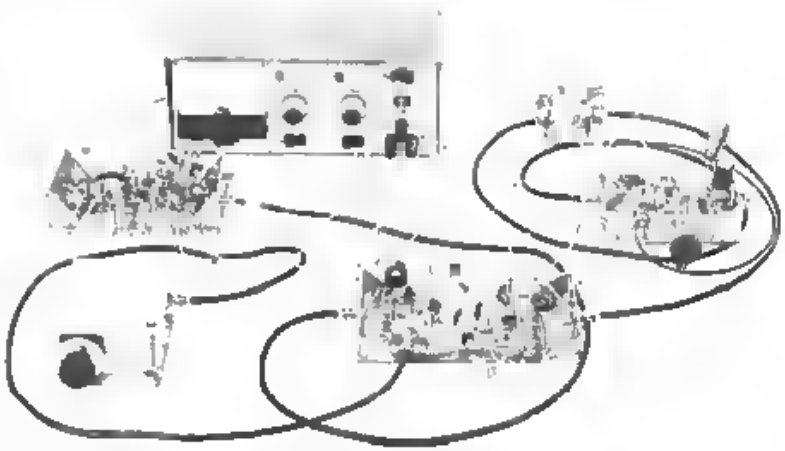


Fig. 7 — The band-imaging receiver prototype. The LO is in the left foreground just behind it, the 10- and 18-MHz RF amplifiers. At center the mixer/filter/IF amplifier module with the detector/AF amplifier module at far right. The smaller module, upper right, is the BFO. Performance was good even though etched circuit boards were not used.

Construction

The receiver was prototyped using point-to-point and "dead bug" modular construction (see Fig. 7). Later, circuit boards were designed and debugged.* You may use either method for building your receiver, with good results. The following construction hints are based on the circuit-board version of the receiver, but much of the information here will be of use to builders using either style of construction.

Parts for this receiver are available from a number of sources. Virtually everything can be obtained from RadioKit, Mouser, Radio Shack, DigiKey and Circuit Specialists. See the parts suppliers list at the end of Chapter 35 for addresses and telephone numbers of these suppliers.

See Fig. 8. The receiver is housed in a Hammond 1590F diecast aluminum box [approximately 7⅛ × 7⅛ × 2⅛ inches]. Threaded standoffs are used to mount all circuit boards except the detector/audio board; spade lugs are used to mount this module vertically. Miniature 50-ohm coaxial cable (RG-174) is used for all RF connections between modules except the LO-mixer line. Here, miniature Teflon® dielectric cable is used because of Teflon's high melting point (see Fig. 9A). RG-174 is also used to connect J3, SIDETONE INPUT, to the detector/audio board. Connections from this board to the AF GAIN and SIDETONE LEVEL controls are made with stranded hookup wire in three colors. This makes for more compact wiring than miniature coax allows and causes no problems with hum or crosstalk. The IF GAIN control and audio output connections are also made in this way. Dc wiring is stranded hookup wire; binding posts are used to bring dc into the receiver.

We recommend that you build, test and install the band-imaging receiver modules in this order: (1) LO; (2) detector/audio and BFO; (3) mixer/filter/IF amplifier; (4)

*Circuit board templates are available from the ARRL Technical Department secretary for $3.00 and a no. 10 s.a.s.e.

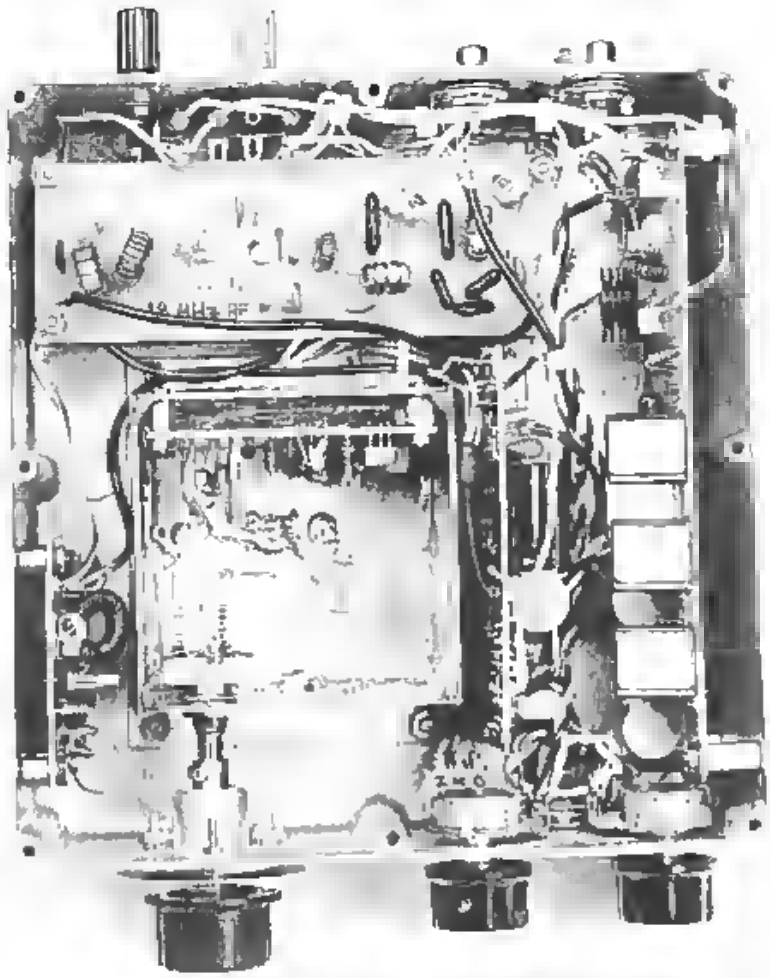


Fig. 8 — The band-imaging receiver installs into a Hammond 1590F diecast aluminum box. In this top view the LO (in its shield box, shown here with cover removed) dominates the layout. The RF amplifier boards are at top, one above the other. The BFO is to the left of the LO, with the detector/audio amplifier board (edge-on) to the right. At far right, the mixer/filter/IF amplifier board. The two capacitors and resistor beneath the mixer/AF board have since been incorporated into the board design. For a similar view of the band-imaging receiver in full color, see the cover of July 1987 QST.

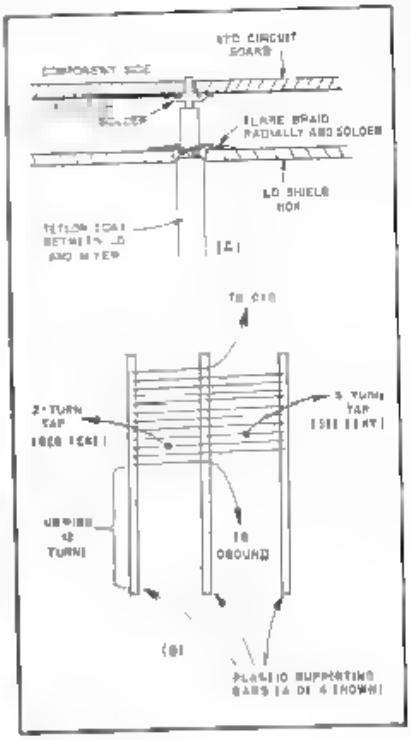


Fig. 9 — The detail drawing at A shows installation of the LO output lead. Miniature Teflon coax is recommended here because its dielectric will not melt during soldering. The shield must be flared and soldered to the *inner* wall of the LO shield box. At B details of the LO inductor, L5, are shown. Taps are made by pushing in adjacent turns of the coil stock. The untinned plastic bars preserve the Q of L5 by holding the Miniductor away from the shield box bottom. Before counting the 11 turns necessary for the coil, unwind enough turns from the uncut Miniductor stock to allow L5 to stand above the shield box bottom by its diameter. The plastic bars are fragile; they may be impregnated with Duco or similar cement of greater strength. The completed L5 is cemented to the shield-box bottom after the LO tuning range has been set.

RF amplifiers. The LO comes first because its installation entails the majority of the metalwork necessary to build the receiver. The sequence allows you to use completed modules as part of your test equipment for the modules later in the sequence.

The LO shield box is made of double-sided copper-clad circuit board. The 10:1 epicyclic reduction drive is a Jackson Bros 5857. Because the sides of the Hammond diecast aluminum box are not perpendicular to the bottom, special construction techniques are needed to ensure that the TUNING capacitor shaft is perpendicular to the front panel of the box. The following construction sequence resulted in a smooth-tuning, no-backlash LO installation in the ARRL lab version of the band-imaging receiver.

1) Mount the 10:1 reduction drive on the front panel.

2) Build the LO shield box (four sides and bottom), soldering only the side and rear pieces into place on the bottom. The front and rear pieces of the shield box must butt the shield box sides as shown in Fig. 8. *Tape* the front side into place. Drill four mounting holes in the shield box bottom plate.

3) Chisel the molded-in printing from the center of the diecast-box bottom to smooth the box floor. Sanding may also be necessary to achieve this.

4) Locate the C9 (TUNING capacitor) mounting hole in the LO shield box front by pushing the box up against the reduction-drive coupler. Size this hole slightly larger than the capacitor mounting bushing. This allows later adjustment of C9's position. Temporarily mount the capacitor in the front side of the shield box and keep this assembly taped to the rest of the box.

5) Place the LO box in the diecast box so that C9 is inserted into the reduction-drive coupler. By feel, be sure that the capacitor is about 1/16-inch short of full insertion into the coupler sleeve. This allows leeway for later adjustment. Mark the diecast box to pass the LO mounting screws through the holes in the shield-box bottom plate. Drill these holes now.

6a) Remove the taped-on front of the shield box. Build and install the LO/buffer circuit board, including the LO output cable and 12-V dc line, in the partially completed shield box. The output cable consists of a 6-inch piece of miniature Teflon coaxial cable (see Fig. 9A). Fig. 9B shows how to prepare LO inductor L5 from a length of B & W Miniductor.

6b) Temporarily install C9, C10, L5 and the 27-pF NP0 LO tuned-circuit capacitor to the LO/buffer board by short leads. Terminate the LO output cable with a 51-ohm resistor. Verify operation of the LO/buffer board by applying dc power and tuning an LO signal on the 14-MHz test receiver. Adjust C9 and C10 as necessary

to bring the signal into your receiver's tuning range. You may need to add or remove fixed capacitors in the LO tuned circuit. Don't spend time now on setting the LO tuning range; that comes later.

6e) Once LO performance has been verified, disconnect C9, C10, L5 and the fixed tuned-circuit capacitor from the LO/buffer board. Install the board into the LO shield box.

7) Install C10 flat to the shield box floor by soldering down its entire tab. Be sure to allow clearance for C9. Bend the C10 stator tab up to clear the box bottom. Connect C10 to the LO/buffer circuit board with tinned no. 18 solid wire. Solder L5 into the circuit; it will be cemented to the box floor later, but do not do this yet.

8) Bolt the LO into the diecast box. Loosely mount C9 in the front side of the LO box. Slide the front LO box nice into place, and at the same time, slide the C9 shaft into the reduction-drive coupler to about 1/16 inch short of full insertion. *Do not* tape the LO box front into place as before.

9) Adjust the reduction drive to bring its coupler worm screws to approximately 10 and 2 o'clock. Set C9 to maximum capacitance without disturbing the reduction drive. Now, with C9 loose in its mounting hole, tighten the worm screws in the reduction-drive coupler.

10) Tighten C9 to the front of the LO shield box.

11) Using the reduction drive, turn C9 back and forth through its range several times to settle the LO box front into position. Depending on how tightly the front is held in place by the LO box sides, you may need to push the sides apart slightly to free the front piece. By eye, the front of the LO box should appear parallel to the front of the diecast box. If all looks well,

12) Solder the front side of the LO box into position.

Final tuning-range and anti-backlash adjustments will be made during alignment and testing of the receiver.

The circuit board placement shown in Fig. 8 works well. Although the position of the LO shield box left little choice as to the placement of the rest of the circuit boards, maximum spacing between the BFO and mixer/filter/IF amplifier boards was decided on *beforehand* to keep the BFO signal out of the IF amplifier circuitry.

Alignment

Test equipment necessary for aligning the band-imaging receiver is a 51-ohm resistor, a receiver capable of CW reception at 14.0-14.2 MHz and 4 MHz ±1 kHz with an S-meter and frequency display resolution of 1 kHz or greater, and a crystal-controlled marker generator capable of providing 0-kHz markers. Equip the coaxial input of the test receiver with a short test cable terminated with alligator clips.

Detector/audio amplifier and BFO. The audio amplifiers require no adjustment. Align the BFO as follows: Without connecting the BFO to the detector, connect a 51-ohm resistor across the secondary of T4. Set C11 (FREQ ADJ) and C12 (OUTPUT TUNE) to mid-range. Apply 12 V dc to the BFO. Set the test receiver for CW reception at 4000 kHz and attach the shield clip of its test cable to the BFO ground foil. Leave the center-conductor clip unconnected. Next, tune to the BFO on the test receiver. Adjust C12 for maximum received signal as indicated by the test receiver S-meter. Adjust C11 to put the BFO at approximately 4000.5 kHz. This completes alignment of the BFO for now. Remove the 51-ohm resistor from the T4 secondary and connect the BFO to the detector with RG-174 cable.

Mixer, filter and IF amplifier. The IF amplifier requires only one adjustment: With 12 V applied to the mixer/filter/IF amplifier board and later stages, adjust C8, IF TUNE, for maximum noise in the speaker or headphones.

Local oscillator tuning range. Connect the LO output cable to the mixer, and apply 12 V to the LO. Tune C9 to the low end of its range, and set the test receiver to 14.060 MHz. Connect the test cable shield clip to the LO box, but leave the center-conductor clip unconnected. Adjust C10 until you hear the LO in the test receiver. Be sure that the unconnected test-cable lead is far enough from the LO tuned circuit to have no effect on the LO frequency. Set the test receiver to 14.155 MHz. Tune the LO upward in frequency until you hear it in the test receiver. With luck, the TUNING capacitor will be nearly at minimum capacitance. Depending on the exact values of the capacitors in the Q4 gate circuitry, however, your LO may not have enough tuning range, requiring that you search downward for it with the test receiver even with the TUNING capacitor at minimum capacitance. If this is so, move the tap on L5 from 2 to 3 turns above ground and readjust the 14.060-MHz band edge with C10. This will increase the tuning range. (You may need to add capacitance in parallel with the 27-pF LO tuned-circuit capacitor to allow C9 to hit the band edge.) C10's tuning range is much larger than that of the TUNING capacitor, so adjust it carefully. With experimentation, you should be able to achieve a TUNING range of between 90 and 150 kHz. Remember that you'll need to make your final band-edge adjustment *after* installation of the LO box cover, be sure to provide a hole in the cover for this purpose, but leave the cover off for now. After you have set the LO tuning range, cement the base of each L5 pillar to the shield box bottom with Duco® (or similar) cement.

RF amplifiers. Install the 10-MHz RF amplifier in the receiver, and solder a 51-ohm resistor from the center conductor of J1 to ground. Connect the crystal calibrator, set for 10-kHz markers, to J1. Set the BAND switch to 10 MHz. Tune in a marker near the center of the tuning range and adjust C3 for maximum signal. Tune in the lowest marker in the range; adjust C5 for maximum signal. Tune in the highest marker in the range, adjust C6 for maximum signal. Because the C5 and C6 adjustments interlock somewhat, repeat them several times for good measure. Now, install the 18 MHz RF amplifier board and repeat this procedure at 18 MHz with C3 (at band center), C5 (at the lowest marker) and C6 (at the highest marker). This completes alignment of the RF amplifier boards.

Anti-backlash adjustment. With luck, the TUNING control will turn freely and require the same input torque across the tuning range. Backlash should be imperceptible throughout the range. If backlash is present, try loosening the reduction-drive coupler screws and tightening them again. Backlash in the ARRL lab version of this receiver was done away with by loosening and retightening the tuning capacitor in its mounting hole, *and* by slipping the TUNING capacitor several degrees to one side in the drive coupling sleeve before retightening the coupler worm screws.

Dial calibration. Calibrate the tuning dial *after* the tuning range has been set *and* any backlash has been ironed out. In the model shown in Fig. 1, calibration of the 10- and 18-MHz TUNING scales differs by the width of a dial marking. The left edge of each mark is used during 18-MHz tuning (18 MHz L); the right edge is used during 10-MHz reception (10 MHz R). Calibration of the full TUNING capacitor rotation (360°; 180° for each band) would make this unnecessary, but one band would tune "backward" relative to the other.

Performance

Measured performance of the band-imaging receiver at 10 MHz: Minimum discernible signal (MDS), −140.5 dBm; two-tone 3rd-order IMD dynamic range (20-kHz spacing), 89.5 dB; blocking dynamic range, 134 dB; image rejection, 74 dB. At 18 MHz, MDS, −140.0 dBm; two-tone 3rd-order IMD dynamic range, (20 kHz spacing), 90.0 dB, blocking dynamic range, 131 dB, image rejection, 82 dB. With a signal tuned in on the 10-MHz band, dropping the receiver three inches to the operating table produced *no discernible shift* in the pitch of the received signal. Maximum audio output was 0.66 W into an 8.2-Ω test load. Current drain at 12 V dc was 95.1 mA with no input signal, 220 mA at maximum audio output.

From June 1976 QST, p 27

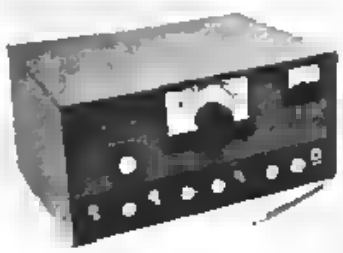
His Eminence—the Receiver

Part 1: No piece of amateur equipment holds greater sway over our communications pastime than the station receiver. Herefrom, let there be dynamic range!

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther, MI 49656

Are you slave to a receiver which unleashes its fury like a many-headed monster in the presence of strong signals? If your receiver shows a will which is most incorrect for an expensive commercial ham-rhack rigging, then you and I are kindred souls! Being a long-term urban dweller amid a barrage of strong local signals, I have had a long-existing need for a receiver with an "uncrunchable" front end. Numerous commercial receivers have been tested at my station, and most provided appallingly dismal performances when W1AW was operating – just two blocks away – and during contests when seldom-heard, nearby stations seemed to pop out of the void to inundate reception. This case for nail biting led to a special-application design which cured my receiver cross modulation, desensitization, and IMD maladies. Some of the design notes offered here should be of interest to amateurs who build station receivers for use in areas of high signal density.

Most of the principles described are well-known ones, but they have been ignored by some designers of imported and US-made receivers. Emphasis seems to be on impressive appearance, high receiver "sensitivity" (whatever is meant by that term), and myriad other features. Along the way somebody forgot the real name of the game . . . *dynamic range*. At least one amateur (W7ZOI) has emphasized the need for careful attention to these matters.[1] Reasonable immunity to front-end collapse is not expensive or difficult to achieve. The results are measured easily in terms of operating convenience and clean reception.



The receiver is built in a homemade aluminum cabinet. A two-tone gray and flat-black paint job has been applied. Black Dymo tape labels are used for identifying the controls in the black area, and gray labels are limited to the gray portion of the front panel. A cut-down Jackson Brothers vernier dial mechanism (two-speed) is used for frequency readout.

Front-End Features

Although the circuit treated here is for a one-band receiver (1.8 to 2.0 MHz), the design procedures are applicable to any amateur band in the hf spectrum. In my case, I employ "drawn converters" to cover 80 through 10 meters. They are founded on the same concepts to be discussed here.

Fig. 1 shows the if amplifier, mixer and post-mixer amplifier. What may seem like excessive elaboration in design is a matter of personal whim, but the features are useful, nevertheless. For example, the two front-end attenuators aren't essential to good performance, but are useful in making accurate measurements (6, 12 or 18 dB) of signal levels during on-the-air experiments with other stations (antennas, amplifiers and such). Also, FL2, a fixed-tuned 1.8- to 2-MHz band pass filter, need not be included if the operator is willing to re-peak the three-pole tracking filter (FL1) when tuning about in the band. The fixed-tuned filter is my preference when the down converters are in use.

The benefits obtained from a highly selective tunable filter like FL1 are seen when strong signals are elsewhere in (or near) the 160-meter band. The rejection characteristics can be seen in Fig. 2. Insertion loss was set at 5 dB in order to narrow the filter response. Part of the circuit was inspired by Sabin's informative *QST* article, where he employed a three-pole Cohn filter with a 4-dB insertion loss.[2] In this example the high-*Q* slug-tuned inductors are isolated in aluminum shields, and the three-section variable capacitor which tunes them is enclosed in a shield made from pc board sections. Bottom coupling is accomplished with small toroidal coils.

Rf amplifier Q1 was added to compensate for the filter loss. It is mismatched intentionally by means of L10 and Lt1 to restrict the gain to 6 dB maximum. Some additional mismatching is seen at L12, and the mixer is overcoupled to the FET tuned output tank to broaden the response (1.8 to 2 MHz). The design tradeoffs do not impair performance. The common-gate rf stage has good dynamic range and IMD characteristics.

The doubly balanced diode-ring mixer (U1) was chosen for its excellent reputation in handling high signal levels, having superb port-to-port signal isolation, and because of its good IMD performance. The module used in this design is a commercial one which contains two broadband transformers and four hot-carrier diodes with matched characteristics. The amateur can build his own mixer assembly in the interest of reduced expense. At the frequencies involved in this example it should not be difficult to obtain performance equal to that of a commercial mixer.

In discussing this circuit with Hayward (W7ZOI), he suggested that I include a diplexer at the mixer output (L13 and the related .002-μF capacitors). The addition was worthwhile, as it provided an improvement in the noise floor and IMD characteristics of the receiver. The diplexer works in combination with matching network L14, a low-pass L-type circuit. The diplexer is a high-pass network which permits the 56-ohm terminating resistor to be seen by the mixer without degrading the 455-kHz i-f. The low-pass portion of the diplexer helps reject all frequencies above 455 kHz so that the post-mixer

[1]Footnotes appear at end of article.

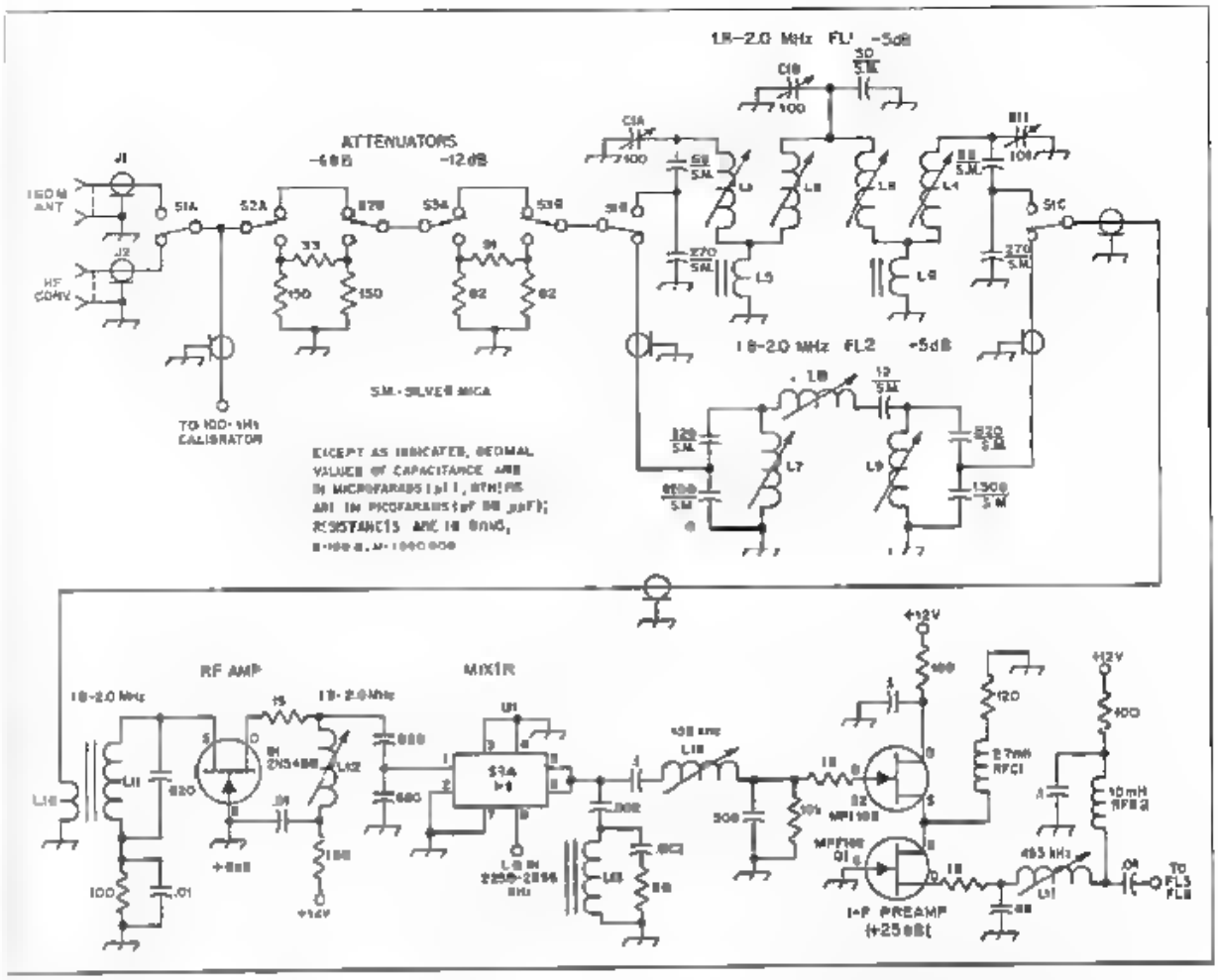


Fig. 1 — Schematic diagram of the receiver front end. Fixed-value capacitors are disk ceramic unless otherwise noted. Resistors are 1/2-W composition. All slug-tuned inductors are contained in individual shield cans which are grounded.

C1 — Three-section variable, 100 pF per section. Model used here obtained as surplus.
J1 — SO-239.
J2 — Phono jack.
L1, L4 — 38 to 58 µH, Q_U of 175 at 1.8 MHz, slug-tuned (J. W. Miller 43A685CB1 in Miller S-74 shield can).
L2, L3 — 95 to 167 µH, Q_U of 175 at 1.8 MHz, slug tuned (J. W. Miller 43A154CB1 in S-74 shield can).
L5, L6 — 1.45-µH toroid inductor, Q_U of 250 at 1.8 MHz.
16 turns No. 26 enam. wire on Amidon T-50-2 toroid.
L7, L9 — 13-µH slug-tuned inductor (J. W. Miller 9052).
L8 — 380-µH slug-tuned inductor (J. W. Miller 9057).
L10 — 16 turns No. 30 enam. wire over L11 winding.
L11 — 45 turns No. 30 enam. wire on Amidon T-50-2 toroid, 8.5 µH.
L12 — 47 µH slug-tuned inductor, Q_U of 50 at 1.8 MHz. (J. W. Miller 9054).
L13 — 8.7-µH toroidal inductor. 42 turns No. 26 enam. wire on Amidon FT-37-61 ferrite core.
L14 — 120- to 280-µH, slug-tuned inductor (J. W. Miller 9056).
L15 — 1.3- to 3.0-mH, slug-tuned inductor (J. W. Miller 9059).
Q1, Q2, Q3 — Motorola JFET.
RFC1 — 2.7-mH miniature choke (J. W. Miller 70F273AT).
RFC2 — 10-mH miniature choke (J. W. Miller 70F102AT).
S1 — Three-pole, two-position phenolic wafer switch.
S2, S3 — Two-pole, double-throw miniature toggle.
U1 — Mini-Circuits Labs SBA-1-1 doubly balanced diode mixer (2913 Quentin Rd., Brooklyn, NY 11229).

For updated supplier addresses see ARRL Parts Supplier List in Chapter 2.

amplifier receives only the desired information. The high-pass section of the diplexer starts rolling off at 1.2 MHz. A reactance of 66 ohms was chosen to permit use of standard-value capacitors in the low-Q network.

A pair of source-coupled JFETs is used in the post-mixer i-f preamplifier. The 10,000-ohm gate resistor of Q2 sets the transformation ratio of the L network at 200:1 (50Ω to 10 kΩ). An L network is used to couple the preamplifier to a diode-switched pair of Collins mechanical filters which have a characteristic impedance of 2000 ohms. The terminations are built into the filters.

Gain distribution to the mixer is held to near unity in the interest of good IMD performance. The preamplifier gain is approximately 25 dB. The choice was made to compensate for the relatively high insertion loss of the mechanical filters — 10 dB. Without the high gain of Q2 and Q3 there would be a deterioration in noise figure.

Local Oscillator

A low noise floor and good stability are essential traits of the local oscillator in a quality receiver. The requirements are met by the circuit of Fig. 3. Within the capabilities of the ARRL lab measuring procedures, it was determined that VFO noise was at least 90 dB below fundamental output. Furthermore, stability at 25°C ambient temperature was such that no drift could be measured from a cold start to a period three hours later. Mechanical stability is excellent. Several sharp blows to the VFO shield box caused no discernable shift in a cw beat note while the 400 Hz i-f filter was

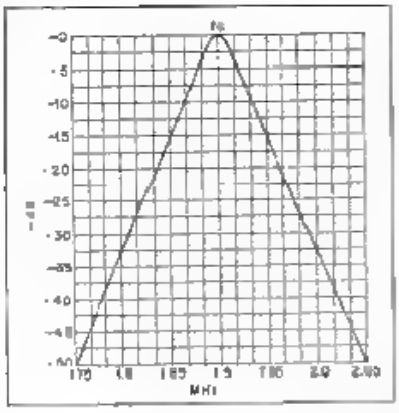


Fig 2 — Response curve of the tunable front-end filter, centered on 1.9 MHz.

actuated VFO amplifier Q14 is designed to provide the recommended +7-dBm mixer injection. Furthermore, the output pr lauk of Q14 is of 50 ohms characteristic impedance. Though not of special significance in this application, the measured harmonic output across 50 ohms is –36 dB at the second order, and –47 dB at the third order.

Filter Modale

In the interest of minimizing leakage between the filter input and output ports, I elected to use diode switching. The advantage of this method is that only dc switching is required, thereby avoiding the occasion for unwanted rf coupling across the contacts and wafers of a mechanical switch. Type 1N914 diodes are used to select FL3 (400-Hz bandwidth) or FL4 (2.5-kHz bandwidth) Reverse bias is applied to the nonconducting diodes. This lessens the possibility of leakage through the switching diodes. Because the Collins filters have a characteristic impedance of 2000 ohms, the output coupling capacitors from each are 120 pF rather than low-reactance .01-µF units, as used at the filter inputs Without the smaller value of capacitance the filters would see the low base impedance of Q4, the post-filter rf amplifier. The result would be one of double termination in this case, leading to a loss in signal level. Additionally, the 120-pF capacitors help to divorce the input capacitance of the amplifie stage. The added capacitance would have to be subtracted from the 350- and 510-pF resonating capacitors at the output ends of the filters

The apparent overall receiver gain is greatest during cw reception, owing to the selectivity of cw filter, FL3. To keep the S-meter readings constant for a given signal level in the ssb and cw modes, R7 has been included in the filter/amplifier module. In the cw mode, R7 is adjusted to bias Q4 for an S-meter reading equal to that obtained in the ssb mode Voltage for the biasing is obtained from the diode switching line during cw reception

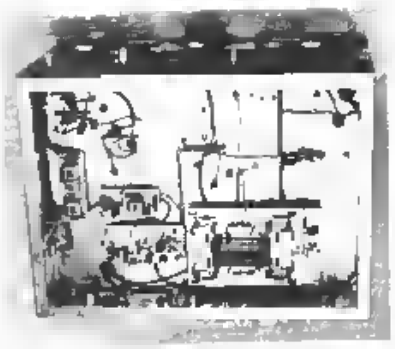
Although a 2N2222A is not a low-noise device, the performance characteristics are suitable for this circuit. A slight improvement in noise figure would probably result from the use of an MPF102, 40673, or low-noise bipolar transistor in that part of the circuit.

Performance Notes

The remainder of the receiver circuit will be discussed in Part 2 of this article However, the reader may want to know just how well *His Eminence* performs, and how the characteristics compare to those of some modern commercial receivers. It seems fitting that the high points be covered in Part 1.

The tuning range of the receiver is 200 kHz. This means that for use with converters the builder will have to satisfy himself with either the cw or the ssb band segment The alternatives are to increase the local oscillator tuning range to 500 kHz, or use a multiplicity of converters to cover the cw and ssb portions of each band. Because 160 meters is my primary band for DXing and casual QSOs during the winter season, the bandspread feature of 200 kHz was adopted.

Some severe lab tests were undertaken with the completed receiver, aimed at learning how "crunchproof" the front end really was. A quarter-wavelength end-fed wire (inverted L) was matched to the receiver 50-ohm input port. The far end of the antenna was situated 3 feet away from the W1AW end-fed Zepp antenna. A pk-pk voltage of 15 was measured across the 50-ohm receiver input jack by means of a Tektronix model 453 scope while W1AW was operating. Now, that's a lot of rf energy! With that high level of rf voltage present, a 10-µV signal was fed into the receiver and spotted 2 kHz away from the W1AW operating frequency. No evidence of cross modula-



Considerable space remains beneath the chassis for the addition of accessory circuits or a set of down converters At the upper left are the adjustment screws for the tunable filter plus the bottom-coupling toroids At the left center is the fixed-tuned front-end filter. To the right is the rf amplifier module A 100-kHz MFJ Enterprises calibrator is seen at the far lower left Immediately to its right is the mixer/amplifier assembly The large board at the lower center contains the i-f filters and post-filter amplifier Most of the amplifier components have been leaked beneath the pc board because of design changes which occurred during development

Fig. 3 — Circuit diagram of the local oscillator differently Resistors are 1/2-W composition mode from pc-board section
C2 — Dandle-bearing variable capacitor, 50 pF.
C3 — Miniature 30-pF air variable.
CR1 — High-speed switching diode, silicon Type 1N914A
L18 — 17- to 41-µH slug-tuned inductor, Q_u at 175 U W. Miller 43A335C91 in Capacitors are disk ceramic unless specified Entire assembly is enclosed in a shield box

Miller 9-74 shield can
L19 — 10- to 18.7-µH slug-tuned pc-board inductor U W Miller 23A155RPC1.
RFC13, RFC14 — Miniature 1-mH rf choke U W Miller 70F103AB.
VR2 — 8.6 V, 1-W Zener diode

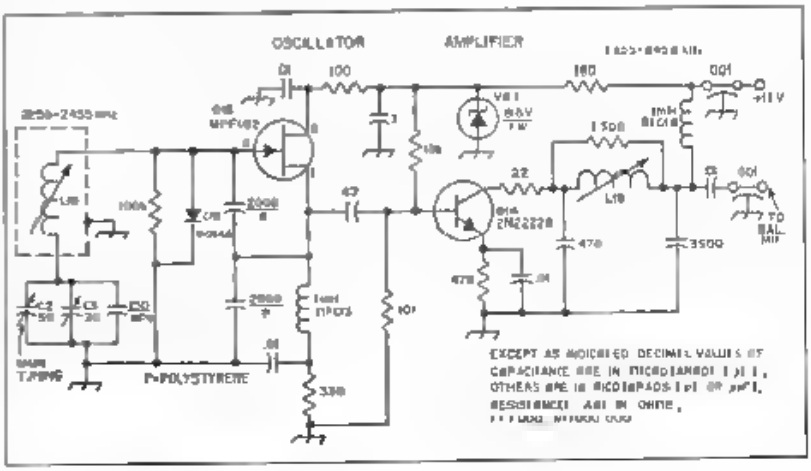
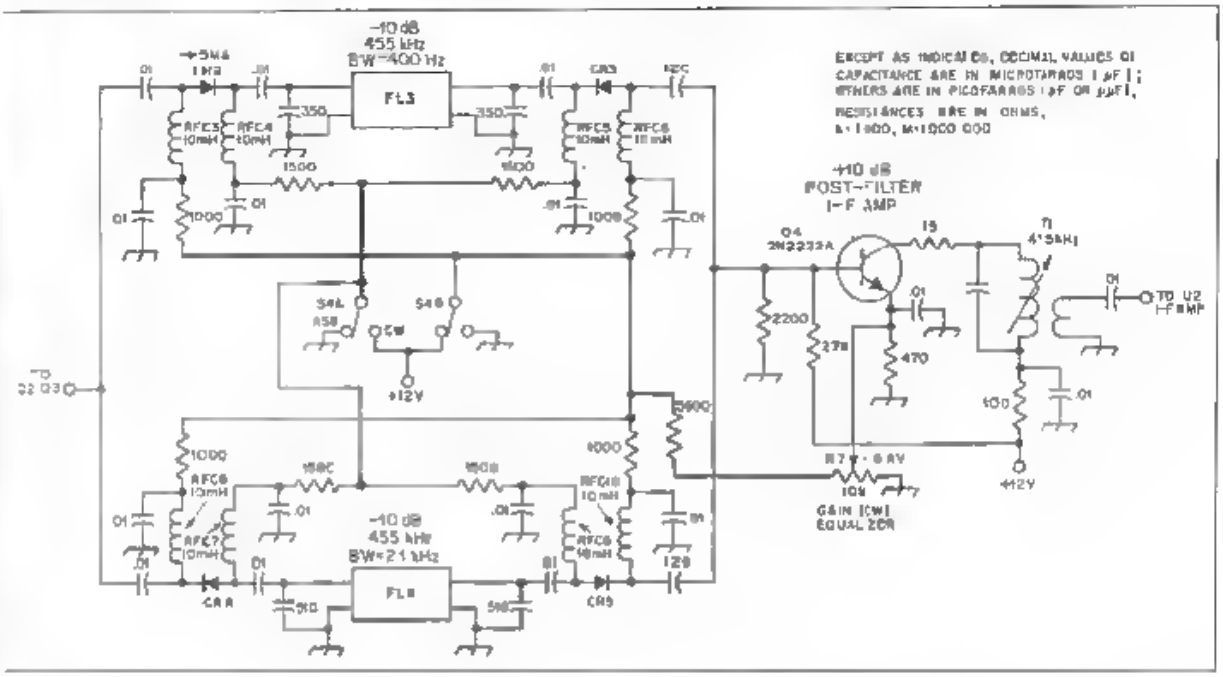


Fig. 4 — Schematic diagram of the filter and i-f post-filter amplifier. Capacitors are disk ceramic. Resistors are 1/2-W composition.

* CR3-CR5, incl. — High-speed silicon switching diode 1N914A.
FL3 — Collins mechanical filter F455FD-04.
FL4 — Collins mechanical filter F455FD-25.
RFC3-RFC10, incl. — 10-mH miniature rf choke (J. W. Miller 70F102A1).
R7 — Pc-board control, 10,000 ohms, linear taper.
S4 — Double-pole, double-throw toggle or wafer.
T1 — Miniature 455-kHz i-f transformer (J. W. Miller 2067, 30,000 to 500 ohms).



tion could be observed, and desensitization of the receiver could not be discerned by ear. The spread from 1.8 to 2 MHz was tuned, and no IM products were heard.

Dynamic range tests were performed in accordance with the Hayward paper in *QST* for July, 1975. Noise floor was −135 dBm, IMD was 95 dB, and 1 dB of blocking occurred at some undetermined point greater than 123 dB above the noise floor. The latter measurement is inconclusive because blocking did not become manifest within the output capability of the model-80 generators used in the ARRL lab. The resultant receiver noise figure at 1.8 MHz is 13 dB which is more than adequate for the high atmospheric noise level on 160 meters.

Table 1 shows measured characteristics for numerous current-model commercial amateur receivers. Brand names can not be listed, but the same test equipment and procedures were used for all checks. It should be kept in mind that the higher the noise-floor figure in −dBm, the better the performance. Similarly, the higher readings for IMD and blocking indicate best performance.

Mute testimony is seen in Table 1. It seems incredible that the three best receivers for IMD and blocking are homemade or modified commercial stock models! It is worth adding that the worst performers are not necessarily the least expensive receivers available. You figure it out, eh? Part 2 of this article will appear in a subsequent issue of *QST*.

Footnotes

[1] Hayward, "A Competition-Grade CW Receiver," *QST* for March and April, 1974.
[2] Salin, "Solid-State Receivers," *QST* for July, 1970.

Table 1

RECEIVER	IMD TWO-TONE DYNAMIC RANGE (dB)	BLOCKING ABOVE NOISE FLOOR (dB)	NOISE FLOOR (−dBm)
W7ZOI Rcvr from *QST* March 1974	95.5	123	141
W1CER Rcvr from *QST* June 1976	95.0	123+	136
Import 1 (modified by ARRL) WA1LNQ	92.5	136	146
Same Rcvr before modification	79.0	110	136
USA Box 1	88.0	116.5	146
USA Box 2	86.0	116	143
USA Box 3	86.0	112	136
USA Box 4	84.0	112	135
USA Box 5	76.0	114	137
USA Box 6 (modified)	83.5	120	141.5
Import 2	79.5	112	139
USA Box 8	79.5	92	123
Import 3	79.0	110	136
USA Box 7	74.5	100	139
USA Box 8	70.0	97	139
Import 4	59.0	102	141
W1CER rcvr. with W7ZOI 20-meter converter attached	88	123+	133

All receivers tested were equipped with 400-, 500-, or 600-Hz i-f filters. Tests were made on 20 meters. Sig spacing = 20 kHz.

From July 1976 QST, p 14.

His Eminence—the Receiver

Part 2: Front end — stay worthy of your vocation with "uncrunchable" distinction! And now the final circuit details.

A receiver i-f system should be capable of providing a specific gain, have an acceptable noise figure, and respond satisfactorily to the applied age. This almost hiomuhe judgment is not as trite as it may stem, for some designers use a haphazard approach to this part of a receiving system. Two of the more serious shortcomings in some designs are poor age (clicky, pumping, or inadequate range) and insufficient i-f gain.

Because of my fringe lassitude and an unwillingness to question past successes, I elected to use a pair of RCA CA3028A ICs in the i-f strip Somewhat greater i-f gain and age range are possible with MC1590G ICs They are the choice of many builders However, the CA3028As, configured as differential amplifiers, will provide approximately 70 dB of gain per pair when operated at 455 kHz This gives an age characteristic from maximum gain to full cutoff which is entirely acceptable for most amateur work

Fig. 5 shows the i-f amplifiers, product detector, and Variable-tuned BFO Transformer coupling is used between U2 and U3, and also between U3 and the product detector The 6800-ohm resistors used across the primaries of T2 and T3 were chosen to force an impedance transformation which the transformers can't by themselves provide Available Miller transformers with a 30,000-ohm primary to 500-ohm secondary characteristic are used U2 and U3 have 10- and 22-ohm series resistors in the signal lines These were added to discourage vhf parasitic oscillations

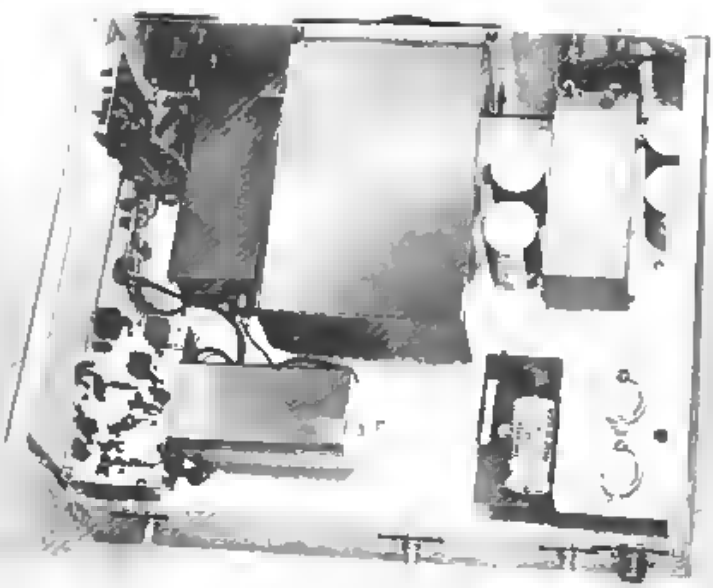
Age is applied to pin 7 of each IC Maximum gain occurs at +9 V, and minimum gain results when the age voltage drops to its low value, +2 V The age is if-derived, with i-f sampling for the age amplifier being done at pin 6 of U3 through a 100-pF blocking capacitor

The 1000-ohm decoupling resistors in the 12-V feed to U2 and U3 drop the operating voltage to +9 This aids stability and reduces i-f system noise The amplifier strip operates with unconditional stability

Product Detector

A quad of 1N914A diodes is used in the product detector. Hot-carrier diodes may be preferred by some, and they may lead to slightly better performance than the silicon units I chose A trifilar broadband toroidal transformer, T4, couples the i-f amplifier to the detector

Top chassis view of the receiver. The R-C active filter and audio preamplifier are built on the pc board at the upper left. To the right of the BFO module in a shield box. The age circuit is seen at the lower left, and to the right is the i-f strip in a shield enclosure. The large shield box at the upper center contains the VFO. To its right is the tunable front-end filter. The three-section variable capacitor is inside the rectangular shield box. The audio amplifier module is seen at the lower right. The small board (mounted vertically) at the left center contains the product detector. Homemade end brackets add mechanical stability between the panel and chassis and serve as a support for the receiver top cover.



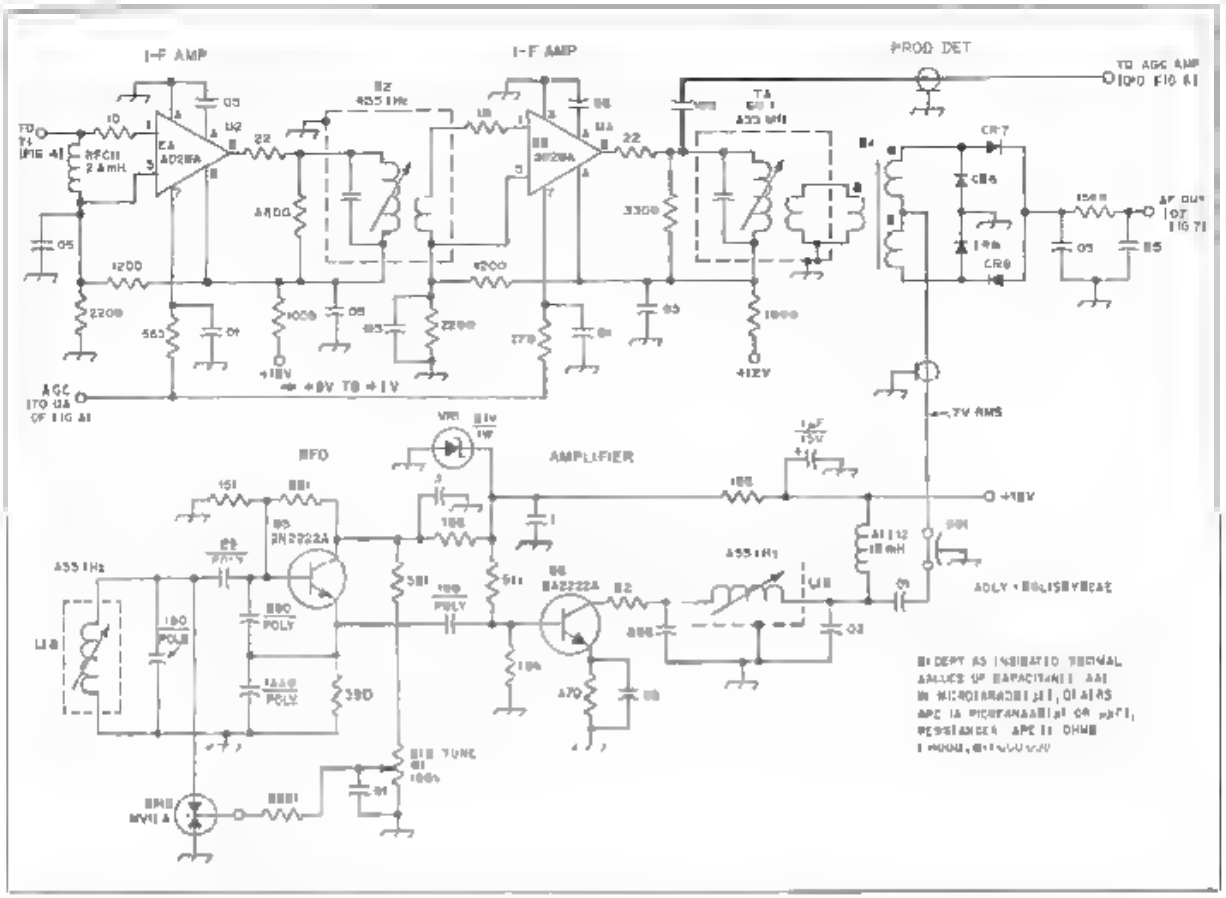


Fig. 5 – Circuit of the i-f amplifier, BFO and product detector. Capacitors are disk ceramic unless noted differently. Fixed-value resistors are 1/2-W composition. Dashed lines show shield enclosures. The BFO and i-f circuits are installed in separate shield boxes. The R-C active filter and af preamplifier are on a common circuit board, which is not shielded.

CR6-CR9, incl. – High-speed silicon, 1N914A or equiv.
CR10 – Motorola MV-104 Varicap tuning diode.
L16 – Nominal 830-µH slug-tuned inductor (J. W. Miller 9057).
L17 – Nominal 60 µH slug-tuned inductor (J. W. Miller 9054).
R1 – 100,000-ohm linear taper composition control (panel mount).
RFC11 – 2.5 mH miniature choke (J. W. Miller 70F253AT).
RFC12 – 0 mH miniature choke (J. W. Miller 70F102AT).
T2, T3 – 455-kHz i-f transformer. See text. (J. W. Miller 2067).
T4 – Trifilar broadband transformer. 15 trifilar turns of No. 26 enam. wire on Amidon T-50-61 toroid core.
U2, U3 – RCA IC.
VR1 – 9.1 V, 1 W Zener diode.

For updated supplier address, see ARRL Parts Supplier List in Chapter 2.

at a 50-ohm impedance level. BFO injection is supplied at 0.7 V rms.

BFO Circuit

In the interest of lowering the cost of this project, a Varicap (CR10 of Fig. 5) is used to control the BFO frequency. Had a conventional system been utilized, three expensive crystals would have been needed to handle upper sideband, lower sideband and cw. The voltage variable capacitor tuning method shown in Fig. 5 is satisfactory if the operator is willing to change the operating frequency of the BFO when changing receive modes. Adjustment is done by means of front panel control R1. Maximum drift with this circuit was measured as 5 Hz from a cold start to a time three hours later. A Motorola MV-104 tuning diode is used at CR10.

To vary the BFO frequency from 453 to 457 kHz, the diode is subjected to various amounts of back bias, applied by means of R1. Regulated voltage (VR1) is applied to the oscillator and tuning diode.

Q6 functions as a Class A BFO amplifier/buffer. It contains a pi-network output circuit and has a 50-ohm output characteristic. The main purpose of the amplifier stage is to increase the BFO injection power without loading down the oscillator.

AGC Circuit

Fig. 6 shows the age amplifier, rectifier, dc source follower, and op-amp difference amplifier. An FET is used at Q10 because it exhibits a high input impedance and will not, therefore, load down the primary of T3 in Fig. 5. Q11 is direct coupled to a pnp transistor, Q11. Assuming that R5 and R2 are treated as a single resistance, Rs, the Q10/Q11 gain is determined as: Gain (dB) = 20 log Rc ÷ Rs. Control R2 has been included as part of Rs to permit adjustment of the age loop gain. Each operator may have a preference in this regard. I have the age set so it is fully actuated at a signal-input level of 10 µV. Age action commences at 0.2 µV (1 dB of gain compression).

Age disabling is effected by removing the operating voltage from Q10 and Q11 by means of S5. Manual if

gain control is made possible by adjusting R3 of Fig 6. Age delay is approximately 1 second. Longer or shorter delay periods can be established by altering the values of the Q14 gate resistor and capacitor. Age amplifier gain is variable from 6 to 40 dB by adjusting R2. The arrangement at Q14 and U4 was adapted from a design by W7ZOI. Age action is smooth, and there is no evidence of clicks on the attack during strong-signal periods. At no time has age "pumping" been observed.

Audio System

A major failing of many receivers is poor-quality audio. For the most part this malady is manifest as cross-over distortion in the at-output amplifier. Moreover, some receivers have marginal audio-power capability for normal room volume when a loudspeaker is used. Some transformerless single-chip audio ICs (0.25- to 2-W class) exhibit a prohibitive distortion characteristic, and this is especially prominent at low signal levels. The unpleasant effect is one of "fuzziness" when listening to low-level signals. Unfortunately, external access to the biasing circuit of such ICs is not typical, owing to the unitized construction of the chips.

Since "sanitary" audio is an important feature of a quality communications receiver, I used a circuit containing discrete devices. The complementary-symmetry output transistors and the op-amp driver are configured in a manner similar to that used by Jung in his *Op Amp Cookbook* published by Howard Sams. Maximum output capability is 3.5 W into an 8-ohm load. An LM-301A driver was chosen because of its low-noise profile. There has been no aural evidence of distortion at any signal level while using the circuit of Fig. 7. The game played in this situation is one of having considerably more audio power available than is ever needed — a rationale used in hi-fi work.

R-C Active CW Filter

A worthwhile improvement in signal-to-noise ratio can be realized during weak-signal reception by employing an R-C active bandpass filter. A two-pole version (FL5) is shown in Fig. 7. A peak frequency of 800 Hz results from the *R* and *C* values given.

The benefits of FL5 are similar to those described by Hayward in his "Competition-Grade CW Receiver" article, which was referenced earlier. He used a second i-f filter (at the i-f strip output) to reduce wide-band noise in the system. The R-C active filter serves in a similar manner, but performs the signal "laundering" at audio rather than at if. The technique has one limitation — monotony in listening to a fixed-frequency beat note, which is dictated by the center frequency of the filter. The R-C filter should be designed to have a peak frequency which matches the cw beat-note frequency preferred by the operator. That is, if the BFO is adjusted to provide an 800-Hz cw note, the center frequency of FL5 should also be 800 Hz.

Experience with FL5 in this receiver has proved in many instances that weak DX signals on 160 meters could be elevated above the noise to a QS copy level, while without the filter solid copy was impossible. It should be stressed that high-*Q* capacitors be used from C4 to C7, inclusive, to ensure a sharp peak response. Polystyrene capacitors satisfy the requirement. To ensure a well-defined (minimum ripple) center frequency, the capacitors should be matched closely in value (5 percent or less). Resistors of 5-percent tolerance should be employed in the circuit, where indicated in Fig. 7.

Summary Comments

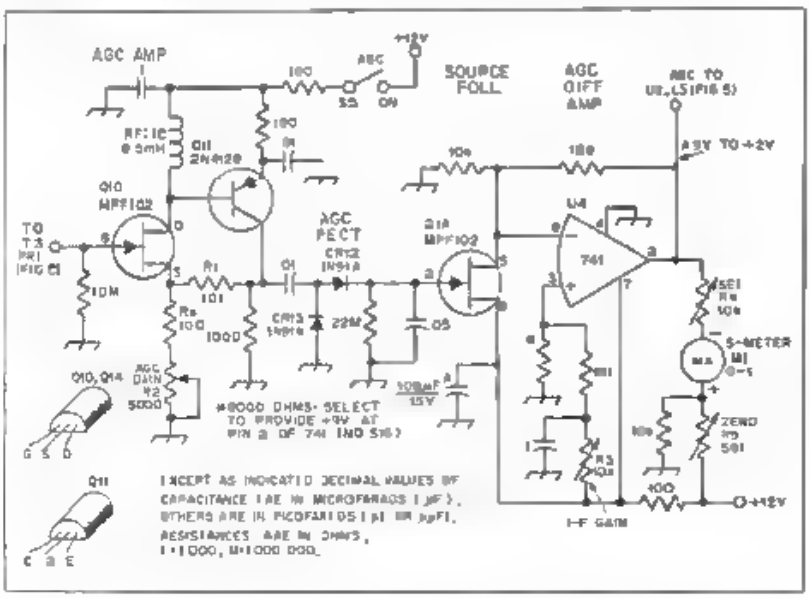
A suitable frequency scheme for some hf-band down converters, plus a circuit for digital frequency display, are given in the receiving chapter of the 1976 *Handbook*. In that example the tunable i-f receiver covers 500 kHz, 1.8 to 2.3 MHz.

The photograph in this article illustrates a modular construction technique. All rf-circuit assemblies are isolated from one another, and from outside energy influences, by means of shield compartments. Signal points are joined (module to module) with RG-174/U subminiature coaxial cable, the shield braids being grounded to the chassis at each end. Feedthrough-type .001-μF capacitors are used at the 12-V entry points of the modules. The foregoing measures help to prevent birdies and unwanted stray rf pickup.

The intent of this paper has been to illustrate some ordinary design principles which can be adopted by those wishing to construct a receiver with wide dynamic range. Some of the ideas offered may inspire modifications to commercial receivers. Because this presentation was not meant as a construction exercise, circuit-board templates are not offered. Most of the pc boards in this prototype have been altered severely during the development pro-

Fig 6 — Schematic diagram of the age system. Capacitors are disc ceramic except when polarity is indicated, which signifies electrolytic. Fixed-value resistors are 1/2-W composition. The module is not enclosed in a shield compartment.

CR12, CR13 — High-speed silicon (1N914A or equiv.)
Q10, Q11, Q14 — Motorola transistor.
R2, R4, R5 — Linear taper composition pc board mount control.
R3 — 10,000-ohm linear-taper control, panel mounted.
RFC15 — 2.5-mH miniature choke (J. W. Miller 70F253AII).
S5 — Single-pole, single-throw toggle.
U4 — Dual-in-line (8-pin 741 op amp.
M1 — 0- to 1-mA meter.



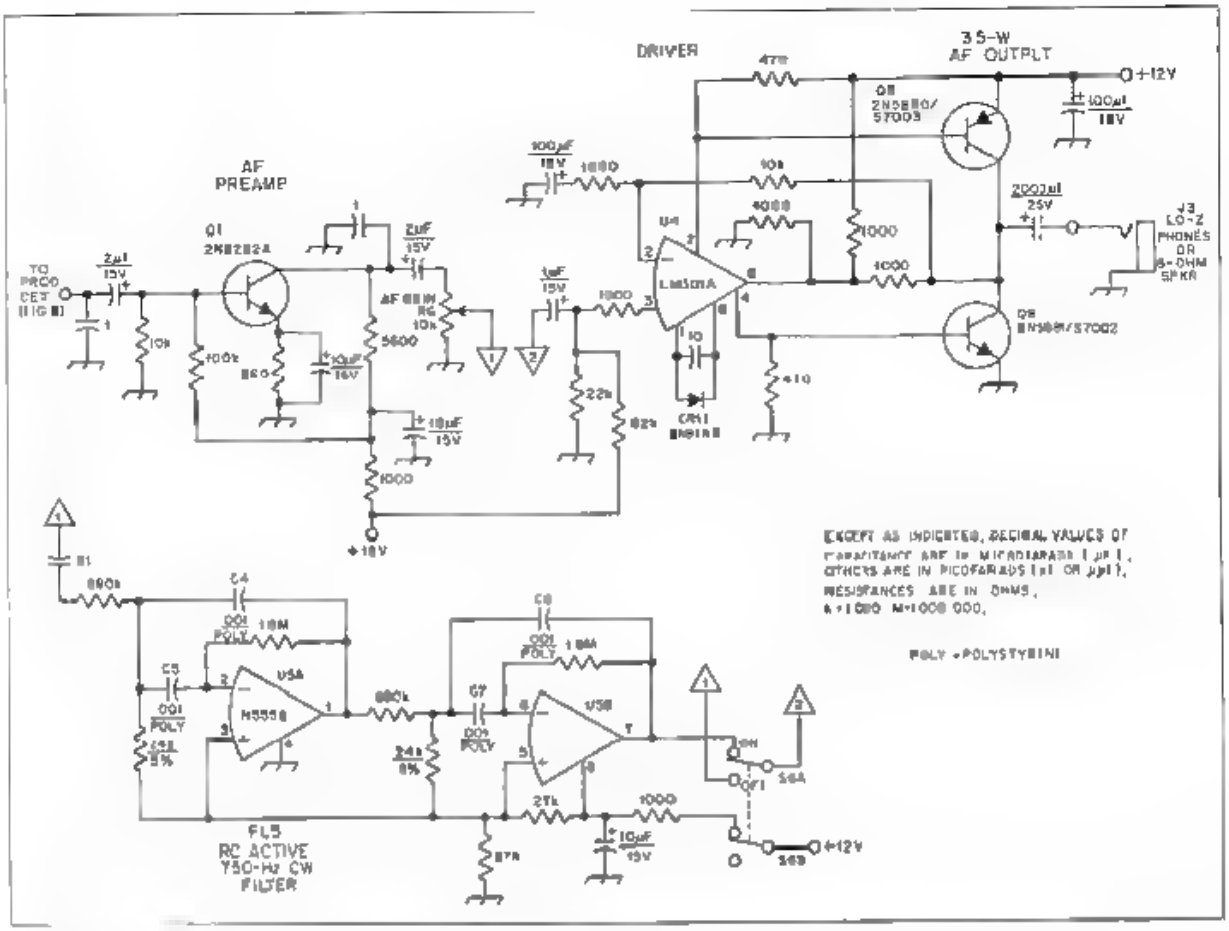


Fig. 7 — Diagram of the audio amplifier and R-C active filter. Capacitors are disk ceramic unless otherwise noted. Polarized capacitors are electrolytic or tantalum. Fixed-value resistors are 1/2 W composition. This circuit is not contained in a shield box. Heat sinks are used with Q8 and Q9.

CR11 — High-speed silicon, 1N914A or equiv.
C4-C7, incl. — See text.
J3 — Phone jack.
R6 — 10,000-ohm audio-taper composition control, panel mounted.
S6 — Double-throw, double-pole toggle.
U4 — National Semiconductor LM-301A IC.
U5 — Signetics N5558 dual op-amp IC.

cess, and numerous components have been tacked on here and there. For this reason, artwork has not been developed.

During several months of daily use, there has never been a case of desensitization or IMD noted, despite my nearness to W1AW and neighboring entities and DXers. *His Eminence* is, indeed, uncrunchable!

From June 1976 QST, p 31:

CER-verters

A family of high-performance hf-band converters for the W1FB (ex-W1CER) 160-meter "His Eminence" receiver.

By Wes Hayward * W7ZOI

If big signals are taking "pot shots" at your collapsing receiver front end, some design changes are probably needed. Here are some guidelines for amateur and professional designers who are interested in improving receiver dynamic range — a sore point with respect to the performance of many modern-day commercial receivers.

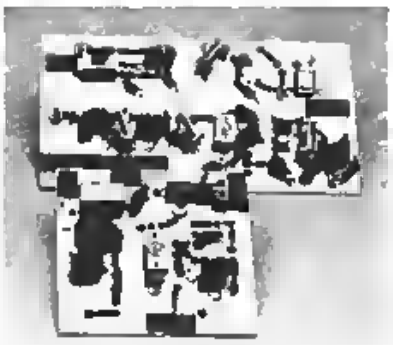
This issue of *QST* contains an article describing some recent receiver efforts at W1CER.[1] That receiver was built to serve two purposes. First, it provided high quality performance on 160 meters. Secondly, and of more significance, it was part of a continuing campaign by W1CER and this writer to develop receivers which meet the classic performance goals of sensitivity, selectivity and stability, while still maintaining a suitable dynamic range.[2,3] As DeMaw pointed out in his two-part article, the amateur can do a much better job than the manufacturer in this regard.

As exciting as the 160-meter band can be, predominant interest is in the hf spectrum. As a result, a group of crystal-controlled converters was needed for the W1CER receiver with an i-f output in the 1.8- to 2-MHz region. Such a family is described here. The primary criterion for their design was to maintain a large dynamic range in a *dual-conversion* system, while still realizing a noise figure that was low enough to be acceptable on the various hf bands.

The information provided to the writer by W1CER was that the minimum discernible signal (MDS), also

[1]Footnotes appear at end of article.

*7700 S.W. Danielle Ave., Beaverton, OR 97005



called the equivalent noise floor of the receiver, was −135 dBm with a 400-Hz bandwidth. Further, the two-tone dynamic range of the receiver was 95 dB.

Information of this type can be related to other more fundamental specifications with a fairly simple set of equations. The noise figure of a receiver is related to the MDS by Eq. 1:

$$\text{MDS(dBm)} = -174\ \text{dBm} + \text{NF(in dB)} + 10\log_{10}B_n \qquad \text{(Eq. 1)}$$

where B_n is the noise bandwidth of the receiver. The noise bandwidth is well-approximated by the 3-dB bandwidth when steep-skirted filters are used, which was the case for the W1CER receiver.

Similarly, the two-tone dynamic range of the receiver is related to the input intercept, P_i, and the MDS by Eq. 2:

$$\text{DR(in dB)} = (2/3)(P_i - \text{MDS}) \qquad \text{(Eq. 2)}$$

where both P_i and MDS are given in dBm. This equation is easily derived from the definition of the intercept concept and the observation that third-order IMD products are proportional to the cube of the strength of the input signals.

A final equation of significance is that which relates the noise factor of two cascaded stages. This relationship, which would apply to a crystal-controlled converter ahead of a receiver, as well as a preamplifier preceding a receiver, is given in Eq. 3:

$$F_{net} = F_1 + \frac{F_2 - 1}{G_1} \qquad \text{(Eq. 3)}$$

In this equation, F_1 and F_2 are noise factors which are algebraic ratios. Noise figure is just the decibel equivalent of this factor. G_1 is the gain of the first stage, again as an algebraic ratio. F_{net} is the noise factor of the combination of a given receiver with a preamplifier or converter with noise factor F_1, and gain, G_1.

From Eqs. 1 and 2, it may be shown that the W1CER receiver had a noise figure of 13 dB and an input intercept of +7.5 dBm. Eq. 3 may be used to infer the overall noise figure when various converter noise figures and gains are considered. The input intercept of a combination will merely be the input intercept of the basic receiver less the gain of the converter. This assumes that the converter is strong enough that minimal IMD occurs within the converter when compared with the following receiver. This implies explicitly that the output intercept of the converter should be much larger than the input intercept of the following receiver.

Converter Designs

After a bit of number "crunching" with the foregoing equations, it was concluded that the converters should have a net gain of about 10 dB and an output intercept of approximately +17 dBm or higher. For work on the bands up through 14 MHz a noise figure of 13 to 16 dB was deemed acceptable. On the higher bands some compromise in

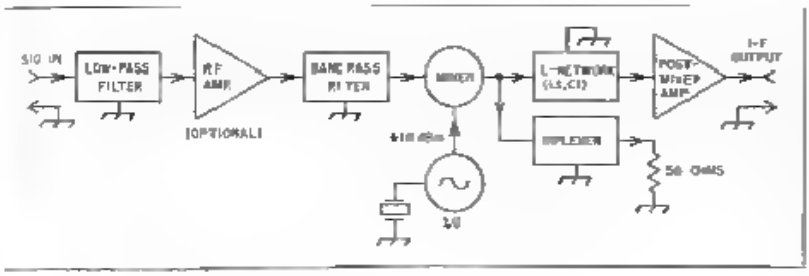


Fig 1 — Block diagram of the CER-verters.

dynamic range would be tolerable in order to achieve lower noise figures. In studying the available circuit combinations it was decided to base the front end of the converters on a diode-ring mixer. The mixer would be preceded by a band-pass preselector filter and followed with a diplexer and a dual-gate MOSFET amplifier at 1.9 MHz. A block diagram of the system is shown in Fig 1.

The original intention was to construct separate converters for each band, 80 through 10 meters. However, after reviewing the design requirements, this was found to be redundant. Diode-ring mixers are inherently broadband and do not require tuned circuits. Further, the post-mixer amplifier would be identical for all of the bands. Only the final-end preselector network and local oscillator need be changed between bands. The final configuration chosen was to use a master board which contained the diode-ring mixer and a post-mixer amplifier. A family of boards was then constructed, each containing a suitable local oscillator and the preselector network for the band of interest.

Mixer and Post-Amplifier Board

The circuit for the mixer and the dual-gate MOSFET amplifier is shown in Fig. 2. There are a few departures from the standard in this design. First, a diplexer is used between the mixer and the "post-amp." This network serves a number of purposes. First, the inductor (L1) and capacitor (C1) driving the FET form an L network which provides an impedance transformation to the gate of the amplifier. A 2200-ohm resistor at the gate assures a termination, causing the mixer to see 50 ohms in the 1.9-MHz frequency range. The other part of the diplexer (C2, C3 and L2) is a high-pass filter designed for a cutoff frequency of 5 MHz. This filter provides a constant i-f termination for the diode ring at virtually *all* frequencies. This is important if the IMD properties of the diode-ring mixer are to be preserved. Such a mixer will create sum-and-difference frequencies from the LO and if inputs. The difference frequency is used to drive the WICER receiver. However, a termination must also be provided for the sum frequency.

In order to simplify the band switching, +12 volts dc is supplied through the local oscillator port of the mixer. This is realized with an rf choke and suitable capacitors.

The output of the amplifier was designed for broadband performance. In order to obtain large bandwidth, the output transformer (T1) was wound on a high-permeability ferrite toroid. A powdered-iron core should not be used for this transformer. Indeed, it was found that a ferrite core with a permeability of 125 was not suitable in this position. Much better bandwidth and impedance matching was obtained with the core specified which has a permeability of 2000. The 2200-ohm resistor in the drain circuit ensures that the output impedance presented by the amplifier is close to 50 ohms. This is important in order to keep the input filters of the WICER receiver terminated properly.

A ferrite bead is used on gate 2 of the amplifier. This may not be necessary in some cases. However, it was included to lessen the possibility of uhf oscillations occurring within the amplifier.[4] A Fairchild FT-0601 or RCA 40673 dual-gate MOSFET can be used at Q1.

Front-End Sections

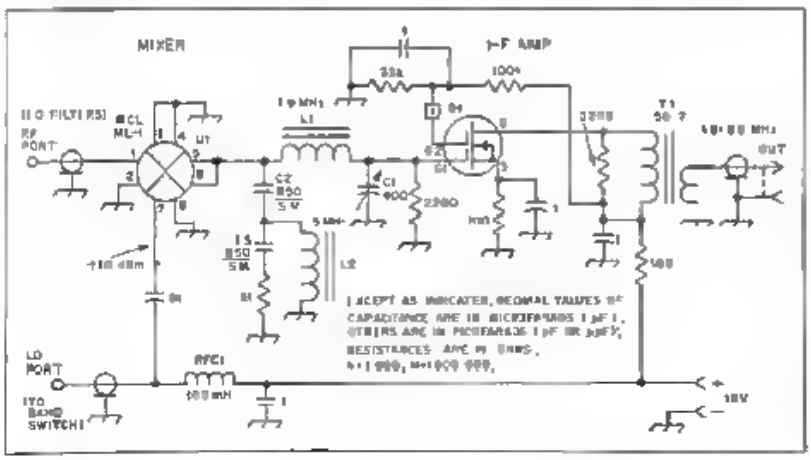
Shown in Fig. 3 is the circuit used as the front end for each of the lower-input bands (3.5-3.7, 7.0-7.2 and 14.0-14.2 MHz). Component values are given in Tables 1 and 2.

The local oscillator for each of the converters uses a bipolar transistor and is designed to provide an output from +10 to +13 dBm. This level of LO injection was found to be near optimum for the diode-ring mixer that was used.

The preselector filters are fairly elaborate. However, the results are well worth the extra expense and effort. Precision-filter-synthesis methods were used to write a computer program for design of the band-pass filters. The coils were wound prior to filter design. Then unloaded *Q* values were measured with a laboratory *Q* meter, and the results were then inserted into the program in order to arrive at the capacitor values. All band-pass filters were designed for a three-pole Butterworth response.

One problem with multisection filters using capacitors as coupling elements between the resonators is that the stopband attenuation may degrade in the vhf spectrum. This is due to slight amounts of lead inductance in the tuning capacitors, and the fact that the capacitive-intersection coupling method degenerates toward a high-pass filter

Fig 2 — Schematic diagram of the master mixer and amplifier circuit. Fixed-value capacitors are disk ceramic unless noted otherwise. Resistors are 1/2-W composition. See Tables 1 and 2 for component values not marked. U1 is a Mini-Circuits Lab ML-1 doubly balanced diode mixer.



response away from the passband. In order to suppress these responses, should they occur, a 5-pole low-pass filter is included at the antenna terminal.

Two methods were used for evaluation of the filter design. First, after initial calculation of the component values, a computer program was used to determine the frequency response of the filters over a wide range. In this analysis, resistors were placed in the circuit to simulate the distortion effects caused by the losses in the cores.[5]

After the filters were built and aligned in the home shop, they were checked with laboratory instrumentation. In this case a Tektronix 7L13 spectrum analyzer and TR-502 tracking generator were used. The measured results around the passband corresponded very well with the computer simulation (which is always encouraging to see). The stop-band attenuation was measured, with one exception, to be over 100 dB for all three filters evaluated. The exception was for the 80-meter filter. At about 70 MHz the attenuation degraded to roughly 95 dB, but returned to the better values at frequencies up through 200 MHz.

One of the reasons a Butterworth response was chosen was that this filter shape is aligned easily with simple test equipment. Alignment is performed by driving the filter with a 50-ohm signal generator and terminating the output in a sensitive 50-ohm detector. The generator is set at the center frequency of the filter and the variable capacitors are adjusted for a maximum response. Experimentally, it was not found necessary to readjust the filters when the swept instrumentation was available.

The converter for the 15-meter band was built using the circuit in Fig 4. On this band it was felt that a better noise figure might be useful. This was provided by inserting an if amplifier between the low-pass filter and the bandpass circuit. The low-pass circuit was modified. The input section is a symmetrical pi network with a Q of 1. This is followed by a pi network with a Q of 10 and an impedance transformation from 50 to 2000 ohms. A 3300-ohm resistor is used in the drain circuit to assure proper termination of the bandpass filter. In the unit built, the drain was attached directly to the hot end of the resonator (L10). However, it would be desirable to reduce the gain somewhat. This would be realized easily by tapping the drain down on the tuned circuit. The terminating resistor should remain across L10.

One problem that the builder may encounter is in obtaining capacitors for the coupling elements between resonator sections of the filter. These values are critical and should not be changed carefully. However, the capacitors may be replaced by a more complicated equivalent network. The basis of this equivalent circuit is to replace a derived capacitor with a series combination of two capacitors with a value which is more than twice the original value. A third capacitor is then placed from the junction of the series capacitors to ground. This configuration is shown in Fig. 5 along with the equations for picking the proper values. As an example, consider the 14-MHz filter, where 3.3-pF coupling capacitors are used. This single capacitor could be replaced with three 10-pF capacitors.

Those building the converter for 80 meters may wish to cover also the 75-meter phone band. While the filter shown could probably be realigned for a range about 100 kHz higher, the shape of the filter would no doubt deteriorate if it were moved further. A better approach would be to change the value of the inductors. Proper results should be obtained by reducing the coils from 35 to 32 turns, keeping all capacitor values the same. A 5.8-MHz crystal would be required for tuning the range from 4.0 to 3.8 MHz.

Table 1

BAND (MHz)	*L3, L4, L8 (TURNS-CORE)*	*L9 (TURNS-CORE)*	*L5, L6, L7 L10, L11, L12 (TURNS CORE)*	*T2, T3 (TURNS-CORE)*
3.5 to 3.7	19, No. 22 T50-2	none	35, No 26 T68-2	25, No 24 T50-2 2 t link
7.0 to 7.2	15, No. 22 T50-2	none	20, No 22 T68-6	25 No 24 T50-2 2-t link
14 to 14.2	12, No 22 T50-6	none	12 No 22 T68-6	21, No 24 T50-6 3 t link
21 to 21.2	10, No 22 T50-6	21, No 22 T50-6	10 No 22 T50-6	19 No 24 T50-6 2 t, link

Coil and transformer data. Toroid cores are Amidon Assoc. powdered-iron type. Y1, Y2, Y3 and Y4 for 3.5 through 21 MHz, respectively, are 5.5, 5.2, 12.2 and 19.2 MHz (International Crystal Co. type GP, 30-pF load capacitance.)

Additional Design Notes

The reader should note that the tuning will be "backwards" for the 80-meter band. This was done for two reasons. First, difficulty was encountered in making the oscillator shown operate properly with the 1.7-MHz crystal that was tried. Of greater significance was the fact that the mixer balance was not especially good at this frequency. As a result, a strong 1.7-MHz signal would have appeared at the input to the post-mixer amplifier. This could have resulted in IMD products. Furthermore, for the 75 meter band the crystal would have been at 2.0 MHz. If low-side injection were used this would have

Fig 3 — Diagram of the filter and crystal oscillator used on 20, 40 and 80 meters. Numbered fixed-value capacitors are silver mica. Resistors are 1/2-W composition. See Table 1 and 2 for coil values.

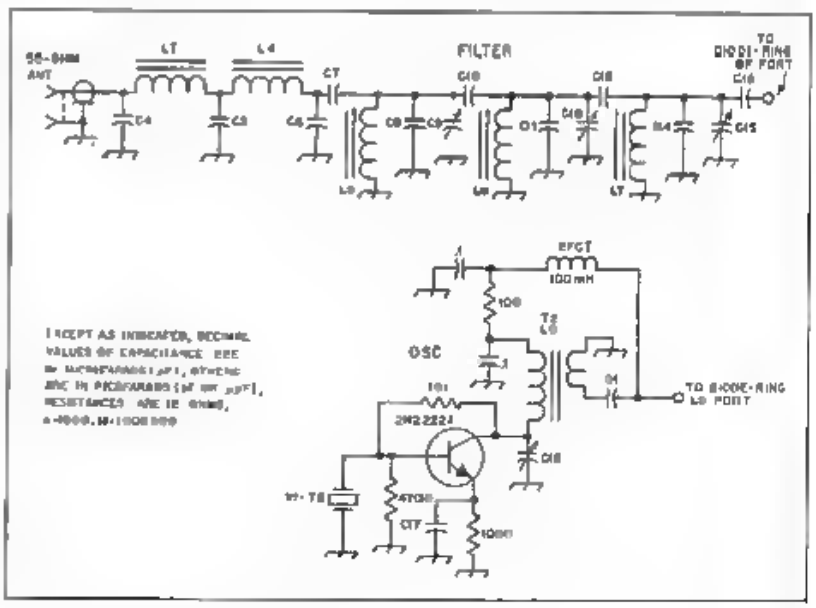


Table 2

BAND (MHz)	C4 C6 C13 (pF)	C5, C20 (pF)	C7 (pF)	C8 (pF)	C9, C12 C15 (pF)	C10 (pF)	C11 (pF)	C13 (pF)	C14 (pF)	C16 (pF)	C17, C31 (pF)	C18, C32 (pF)	C21 (pF)
3.5 to 3.7	790	1580	130	—	90 to 400	12	—	10	—	91	100	400	—
7.0 to 7.2	450	890	43	—	90 to 400	4.7	—	4.7	—	62	100	400	—
14 to 14.2	220	450	33	90	20 to 90	3.3	90	3.3	90	22	47	20 to 90	—
21 to 21.2	150	300, 345	—	51	20 to 90	1.2	51	1.2	51	12	47	20 to 90	20 to 90

Fixed-value and trimmer capacitors. Fixed-value capacitors are silver-mica or similar high-*Q*, stable types. Trimmers are mica compression type. See text for obtaining precise non-standard fixed-capacitance values.

placed a strong signal within the tuning range of the main receiver. If it is desirable that all hf bands tune in the same direction, the builder should pick high-side crystals for all of the bands.

The approach used for the 15-meter converter in order to obtain low-noise performance could also be applied to the 10- and 6-meter bands. The image rejection might be a little poor with such a low i-f in the 6-meter case.

Another revision that many builders may consider would be the construction of a high-performance 80-meter receiver with converters for the higher bands. The converters described would be suitable for this situation. The crystal frequencies would change accordingly. The diplexer between the diode mixer and the "post amp" should be redesigned. This could be done easily by halving the inductance and capacitance values used in the diplexer circuit. The broadband output circuit in the diagram of Q1 should work equally well at 3.5 MHz. The 15- and 20-meter band-pass filters were designed with enough bandwidth to cover the total band. This was done in order to keep the insertion losses at a reasonable level. A slightly wider filter would be required for the total 40-meter band.

The converters are built on rather large circuit boards. This was done in order to ensure a reasonable level of stopband rejection in the filters, and to ease construction. Those interested in a more compact format should consider the inclusion of shields between the sections of the input band-pass filter and between the filter circuitry and the corresponding oscillators. It is fun to build miniature equipment when there is a good need for small size. However, for high-performance home-station equipment, where considerable experimentation may be required, a larger format is often desirable.

Because the pc boards shown in the photograph are quite large, the builder will probably elect to lay the circuits out for a more compact format. For this reason there are no pc-board templates and layouts available.

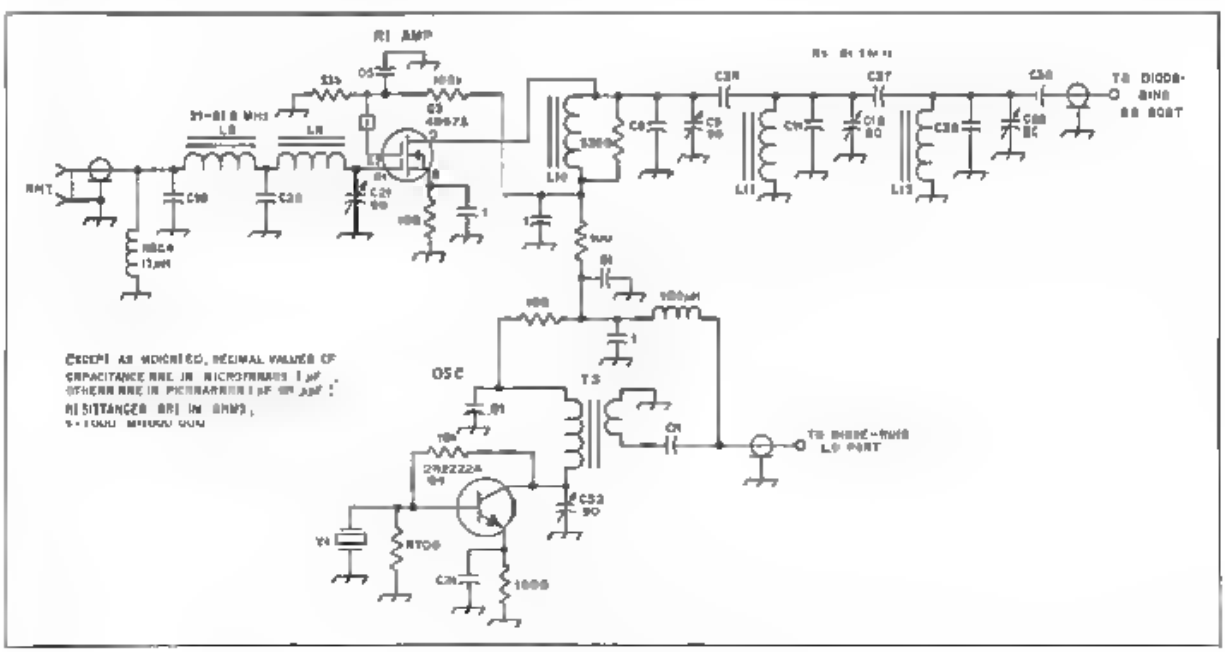
Great care should be taken when the front-end sections are band switched. Shielding between switch wafers should have over 100 dB of isolation. Diode switching is not recommended unless the builder has equipment to evaluate the affects on IMD.

Evaluation and Performance

This project was in some ways quite frustrating, for the W1CER receiver was 3000 miles away. This is the first piece of receiving gear that the writer has built which could not initially be evaluated "by ear." However, a suitable substitute was available for laboratory evaluation. This was a Tektronix 7L5 Spectrum Analyzer. This instrument was extremely convenient to use for this purpose, since it is synthesized with a 250-Hz accuracy, and has resolution down to 10 Hz. The dynamic range is excellent.

The only converter evaluated for IMD was the 14-MHz unit. Two-tone IMD measurements were performed and it was found that the output intercept of the converter was +22 dBm. This is

Fig. 4 — Diagram of the 15-meter front-end circuit. Numbered fixed-value capacitors are silver micas. Resistors are 1/2-W composition. See Tables 1 and 2 for other parts values.



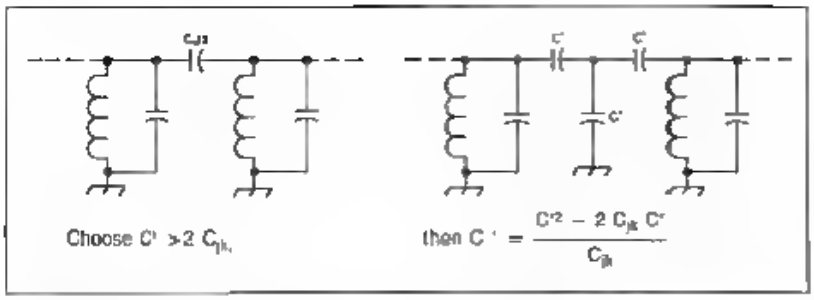


Fig 5 — Method for obtaining filter coupling capacitor of computed values.

more than sufficient for the application, since it greatly exceeds the input intercept of the W CER receiver, +7.5 dBm.

The gain and MDS were measured for all four converters. To remove the effect of the high noise figure of the 7L5 (19 dB, a low-noise MOSFET preamp was built at 1.9 MHz. This unit had a noise figure under 2 dB, allowing meaningful measurement of converter MDS. The signal generator used was an HP-8640B. On the three lower bands, the resultant noise figure of the converters was 12 dB, plus the loss of the input filters. Similarly, the gain of the converter was 12.5 dB, minus the loss of the input filters. It was found that the gain and noise figures could both be improved by removing the 2200-ohm resistor at the gate of Q1. There was a slight reduction in the output intercept, but not enough to cause problems. However, the low-pass part of the diplexer became much sharper in frequency response. This would make a front panel trimmer control necessary.

The 15-meter converter performed differently. The net gain of this unit was 32.5 dB and the noise figure was about 3 dB. This is actually too much sensitivity to be usable at this frequency. It is highly recommended that the builder move the drain tap on the band-pass filter as outlined.

On the basis of the measured results and the published data for the WICER receiver, the system results may be calculated. Shown in Table 3 are the predicted system noise figure, MDS for a 400-Hz bandwidth, input intercept and two-tone dynamic range for the converters operating into the DeMaw receiver. Also shown are the measurements that were obtained for image rejection and i-f feedthrough for the four converters.

Table 3

BAND (METERS)	NFIG (dB)	MDS (dBm)	P_3 (dBm)	DR (dB)	IMAGE REJ. (dB)	I-F FEEDTHROUGH (dB)
80	14 8	−133 2	−2 5	87 1	−90	−110
40	16 3	−131 7	−1 0	87 1	−110	−94
20	16 3	−131 7	−1 0	87 1	−95	−112
15	31	−145	−25 8	80 0	−100	not measured

It is interesting to note that the dynamic range of the system has decreased from 95 dB on 160 meters to 87 dB on the hf bands. This decrease is to be expected in any multiconversion system. Note also that the dynamic range is constant on the three lower bands. This results because the only variation between bands is in the insertion loss of the preselector filters. This difference is the same as would be obtained by adding attenuation in the front end of the receiver. An attenuator will change both the MDS and the input intercept by the same amount, leaving the two-tone dynamic range as a constant of the system. While front-end attenuators are useful accessories for the receiver, they will not improve the dynamic range as it sometimes implied.

A more careful application of attenuation can, however, result in an improved dynamic range. Consider the effect of switching in the 6-dB input attenuator of the WICER receiver, *after* the converters. The input intercept of the 160-meter tunable i-f will now increase to +13.5 dBm and the noise figure will become 19 dB. If the net result is evaluated using the earlier equations, the 20-meter MDS will degrade by only 0.7 dB, but the system input intercept will move up to +5 dBm, leaving a net dynamic range of 90.6 dB. This is a dramatic demonstration of the effect of gain distribution upon dynamic range, especially in multiconversion receivers.

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[4] Belser, "Low Noise 29 MHz Preamplifier for Satellite Reception," *Ham Radio*, Oct, 1975.
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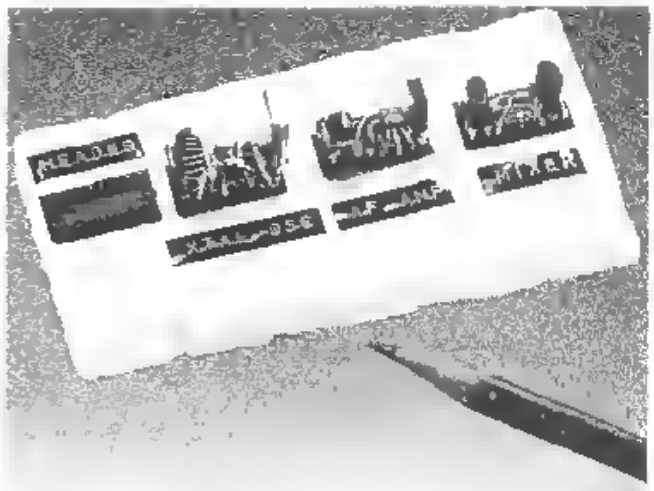
From June 1987 QST, p 30.

Build Your Own MCM ICs



MCMs (mini circuit modules) are fun to lay out and build. With a few IC headers and some patience, you can develop miniature subassemblies that may be used many times.

By Doug DeMaw, W1FB
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Have you considered building your own ICs? The idea is not as ridiculous as it may seem! We must accept the fact that none of us are equipped to construct classic monolithic ICs, wherein the circuit elements are developed on a common piece of silicon (substrate). But it is not mandatory that circuits to be integrated are formed in that manner. With a reasonable amount of imagination and time, it is possible to place your favorite small circuit on a tiny blank DIP header. For lack of a better acronym, let's call these assemblies MCMs (for "mini circuit modules").

You may wonder what the purpose of such an exercise might be. First, we are forced to develop a compact circuit layout, owing to the restricted number of IC-header pins available, plus the small rectangular area of the header. I find that the circuits I have placed on IC headers would typically occupy three times the space on an ordinary PC board. In other words, when there is room to spare, I seem to use it! Miniaturization is beneficial when we wish to build compact gear for portable use, especially for QRP applications.

Another advantage realized from MCMs is that they can be used many times in numerous projects. The same circuits, if built on PC boards, would require complete stripping of the components in order to transplant them on a new PC board. The MCMs can simply be removed from IC sockets and plugged into a socket on some other PC board. This technique should appeal especially to the experimenter or the frugal amateur.

Some Common Circuits as MCMs

Fig 1 shows a compound, direct-coupled audio amplifier that has been built on a 16-pin DIP header.[1] This amplifier has the ability of providing 40 dB of gain, depending on the electrical characteristics of the transistors used. Such an amplifier is suitable for driving a pair of headphones, serving as a mic amplifier or for use ahead of an audio power IC, such as an LM386. Fig 1B shows how the parts are assembled on the header. The heavy outline around the circuit of Fig 1A indicates which components are on the header. C3, for example, is external to the MCM.

Doubly Balanced Mixer MCM

Diode-ring DBMs are available as commercial units in DIP IC packages. Mini-Circuits has some nice units that come in sealed metal packages for direct insertion into IC sockets. These modules are expensive when purchased in single-lot quantities. If one buys 10 or more units, the price becomes more equitable, but few of us want a drawer filled with DBMs that may never be used! So, the MCM approach becomes worth considering for most of our amateur needs.

Fig 2A shows the circuit of a DBM. The diodes should be matched as closely as possible to ensure proper circuit balance. Diode matching may be done by means of a VOM. Sort through your 1N914 or similar small-signal silicon switching diodes and select four that have the same forward-resistance reading (typically between 7 and 12 ohms). Hot-carrier diodes are even better for use in a DBM, and most of them from a given brand and type number are fairly well matched.

Z2 of Fig 2A may be used as a mixer, balanced modulator or product detector. No internal changes are needed, but the exterior circuitry will differ somewhat, depending upon the application. T1 and T2 are miniature broadband transformers. I used two small ferrite balun cores for T1 and T2, but tiny 850 μ toroid cores allow construction of a more compact MCM.

It is essential to connect the T1 and T2 windings as shown. The black dots indicate the polarity of the windings. Local-oscillator injection for this type of mixer is approximately +7 dBm for best IMD performance. Conversion gain (actually a loss) is on the order of −8 dB for a DBM. At frequencies greater than 40 meters, if the DBM is used as a receiver mixer or direct-conversion (D-C) receiver product detector, it is wise to use an RF amplifier between the antenna and the DBM. If not, the receiver noise figure will be too high for weak-signal reception. In fact, an RF amplifier would be an asset even at 7 MHz.

An ideal DBM would be enclosed in a metal case to minimize stray signal pickup. However, there should be no problems with unwanted pickup of RF energy when using the MCM of Fig 2, provided ordinary PC-board layout is employed. In other words, don't place the DBM close to an unshielded oscillator or antenna lead.

Crystal-Oscillator MCM

A simple crystal oscillator is presented in Fig 3A. You may prefer to exclude the crystal, Y1, from the MCM. This will make the module more universal in application. I included the crystal for the purpose of demonstrating the practicality of having Y1 mounted on the IC header. An HC-18/U

[1] IC headers (sometimes called DIP adapter plugs) are available from many surplus electronics parts dealers, including Mouser Electronics, 11433 Woodside Ave, Santee, CA 92071, and ALL Electronics Corp, 905 S Vermont Ave, Box 20406, Los Angeles, CA 90006.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

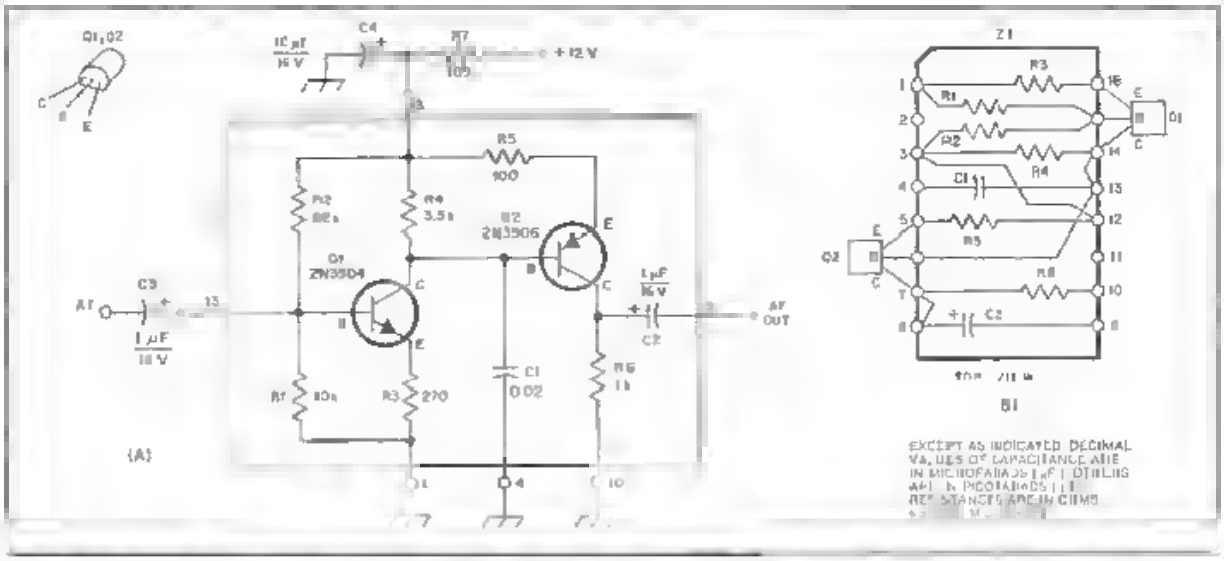


Fig 1—Schematic diagram of a two-stage audio amplifier that can provide up to 40 dB of gain. The drawing at B shows the component layout for the audio MCM.

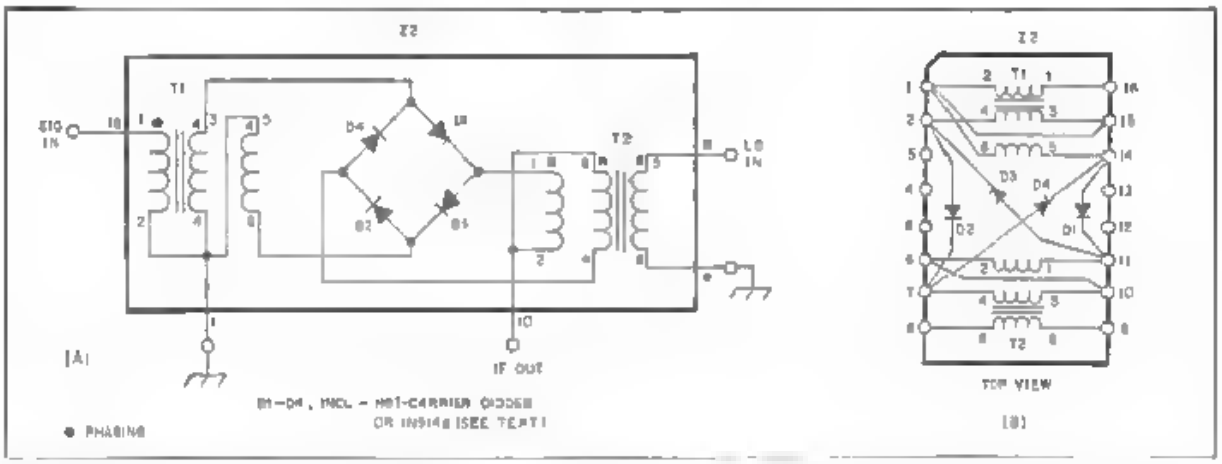


Fig 2—Circuit for a doubly balanced mixer, balanced modulator or product detector. Matched diodes are necessary for best DBM balance. Hot-carrier diodes are recommended for this circuit, but matched 1N914s are suitable. T1 and T2 in my MCM consist of four bifilar turns of no. 30 enameled wire through an Amidon balun core no. BN-43-2402. An FT-23-43 ferrite toroid core may be substituted (smaller) by winding 7 bifilar turns of no. 30 enameled wire on each core.

crystal holder is necessary (small) in order to find room for it on the header. If the crystal will be used outboard from the MCM, you may connect it to pins 1 and 16 (Fig 3B).

The oscillator of Fig 3A is easy to work with. External capacitor C2 is used to control the feedback. It functions in connection with the transistor internal capacitance (C_j) to form a feedback network. C2 should have a capacitive reactance of roughly 200 ohms for most small-signal transistors. This equates to 100 pF for operation at 8 MHz. If you wish to convert this circuit to a VXO, you need only to separate pins 1 and 2 (remove jumper) of the header and place a small inductance (25 µH for 8 MHz) in series with a 75- or 100-pF variable capacitor from pin 1 to ground. Connect the capacitor rotor to ground. This arrangement will provide approximately 6-10 kHz of frequency change.

C3 and L1 are outboard from the MCM. This tuned circuit is resonant at the crystal frequency. L2 is a small link for coupling the circuit to low-impedance loads. For most applications, C3 may be a small trimmer.

Oscillator Buffer/Amplifier MCM

It is seldom necessary to use buffering after a crystal oscillator, since changes in load (reactance changes) seldom cause oscillator pulling. VXOs, on the other hand, may be prone to pulling effects from load changes, and a buffer is useful in that case. VFOs are affected significantly by load changes. Therefore, it is wise to include a buffer or buffer/amplifier after a VFO.

Output coupling from the oscillator should be as light as possible to minimize pulling. Light coupling (C4 of Fig 5A) causes reduced power output from the VFO. As a result of this condition, it is advisable to amplify the VFO output energy to compensate for the power loss. Fig 4A shows a suitable buffer/amplifier circuit that will fit on a 16-pin IC header. Q1 is purely a buffer, and has a gain of 0.9

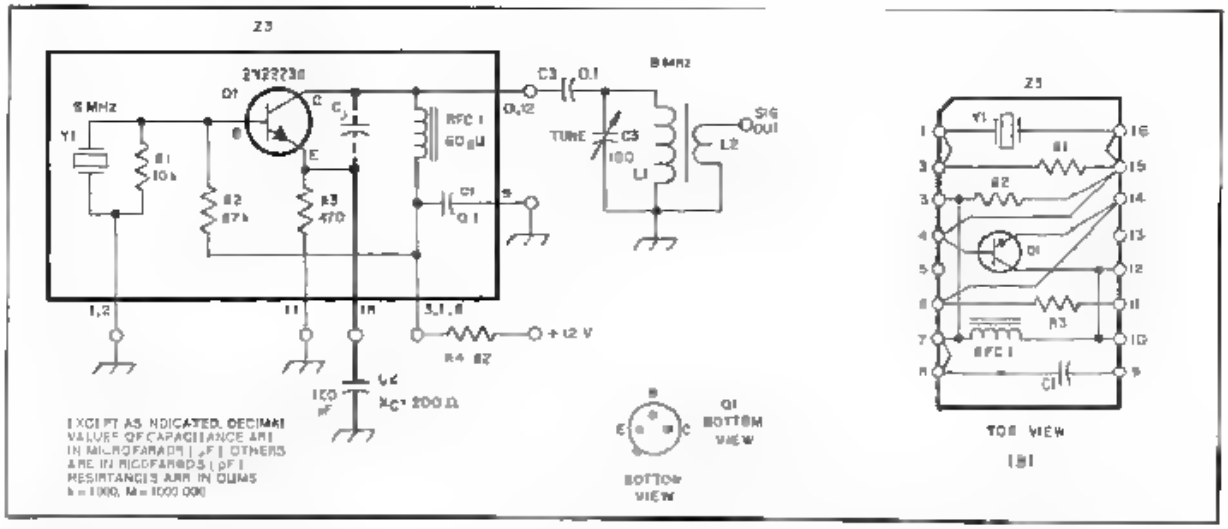


Fig 3—Example (A) of a crystal oscillator that can be built on an IC header. The heavy black line indicates the MCM boundary. All other parts are external to the MCM. L1 is a 5.5-µH inductor (33 turns of no. 28 enameled wire on an Amidon T50-2 core). L2 consists of 6 turns of no. 28 wire. The MCM layout is given at B.

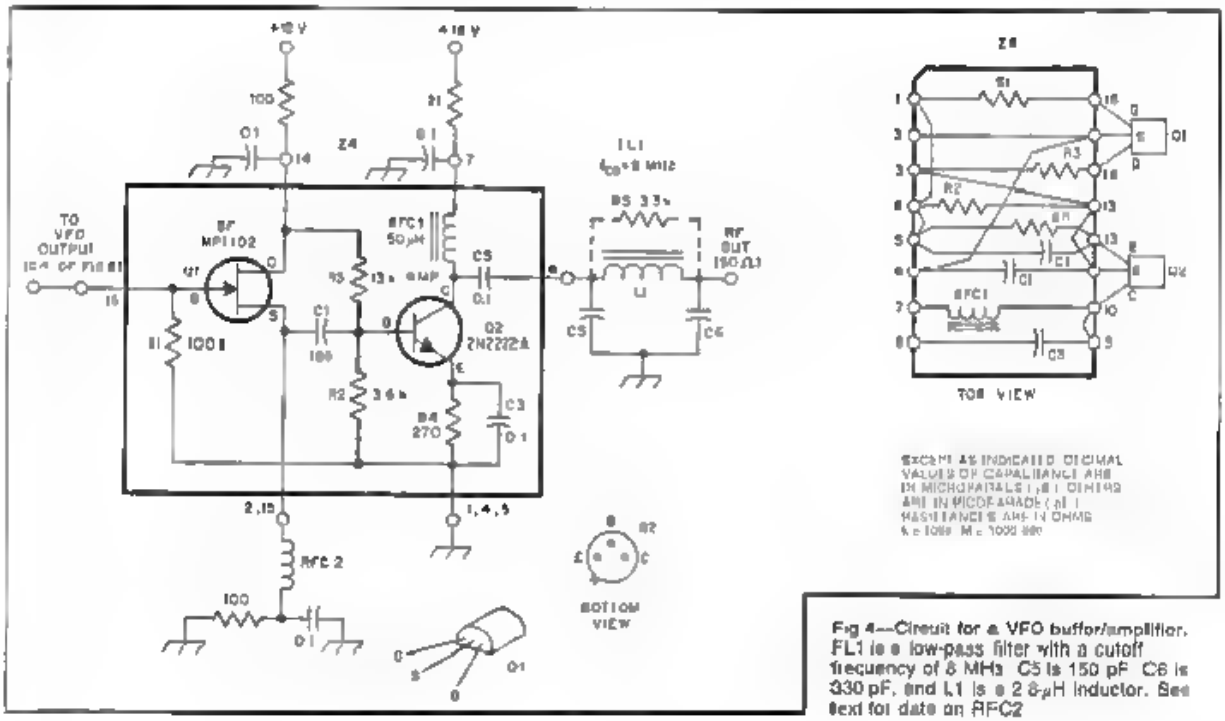


Fig 4—Circuit for a VFO buffer/amplifier. FL1 is a low-pass filter with a cutoff frequency of 8 MHz. C5 is 150 pF. C6 is 330 pF, and L1 is a 2.8-µH inductor. See text for data on RFC2.

typically. This is par for a source to lower. It helps to isolate the VFO from Q2 and the circuits that follow Q2. Amplifier Q2 builds up the VFO energy to a level that is suitable for most circuits with which a VFO is used.

External to Z4 of Fig 4A is a pi network that serves as a matching circuit between the collector of Q2 and a 50-ohm load. This network also serves as a harmonic filter. R5 may be added to increase the loaded bandwidth of the pi network. This may be helpful when the VFO covers a fairly wide frequency range.

RFC2 of Fig 4A is chosen to yield a broad frequency-response peak at the VFO operating frequency. You may assume approximately 10 pF of stray parallel circuit capacitance for RFC2. Thus, for 40-meter operation we will require a 90-µH inductor for RFC2. Should Q2 become unstable, place a 1-kilohm resistor in parallel with RFC1. The addition of R5 will also aid stability in stubborn cases.

Colpitts VFO MCM

The VFO of Fig 5A uses electronic tuning. VVC (voltage variable capacitor) diodes are specified for D1 and D2. This eliminates the need to locate expensive and scarce miniature variable capacitors for tuning VFOs. It is proper to state that long-term VFO drift may be increased through

the use of tuning diodes, as opposed to air variable capacitors. This is because two additional semiconductor junctions have been introduced to the oscillator circuit. Junction capacitance changes with temperature. Normally, the small degradation in frequency stability is acceptable for amateur work. Tuning is done by means of a panel mounted potentiometer (R3). Smooth tuning will result if a 10-turn Helipot® and dial are used, or if a standard potentiometer is used with a vernier drive. The values for R2 and R4 are chosen for the frequency coverage desired, and this will depend upon the type of VVC diodes used for D1 and D2. VVC diodes come in many capacitance ranges. I have suggested for this circuit a pair of diodes that will provide a fairly linear capacitance swing of 10 to 30 pF.

Outboard components C5, C6 and L1 are chosen for the VFO operating frequency. NP0 capacitors are recommended for best overall frequency stability. C1, C2 and C3, internal to the MCM, are also NP0 ceramic capacitors. The smaller 50-V types are preferred in the interest of fitting them on the IC header.

Q1 of Fig 5A may be any high-transconductance JFET, such as a 2N4416. A dual-gate MOSFET may be substituted by tying gates 1 and 2 together and treating the device like a JFET.

Place a shield compartment around the VFO MCM site on the main PC board of a receiver or transmitter. This will help prevent stray RF energy from entering the VFO circuit and causing frequency instability. The small shield compartment may be fashioned from PC-board sections or from flashing copper.

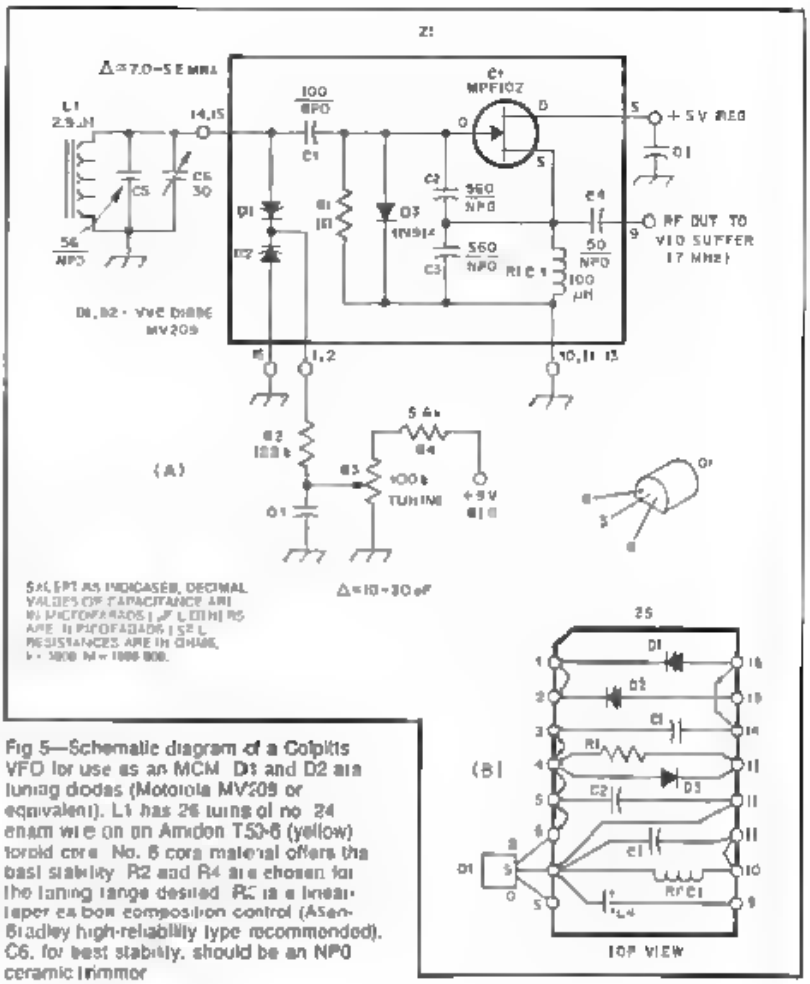


Fig 5—Schematic diagram of a Colpitts VFO for use as an MCM. D1 and D2 are tuning diodes (Motorola MV209 or equivalent). L1 has 26 turns of no. 24 enam wire on an Amidon T50-6 (yellow) toroid core. No. 6 core material offers the best stability. R2 and R4 are chosen for the tuning range desired. R3 is a linear-taper carbon composition control (Allen-Bradley high-reliability type recommended). C6, for best stability, should be an NP0 ceramic trimmer.

MCM Practical Considerations

Miniature equipment is not easy to build, and MCMs certainly fit this description. You will need patience during the assembly procedure, but your skill and speed of construction will increase with practice.

Plug the IC header into an IC socket *before* commencing the MCM assembly. This will prevent the pins of the header from becoming bent or broken. I use an IC socket that has its pins mashed flat against the bottom of the socket. This allows the socket to lie flat on the bench during assembly. A "third hand" type of soldering fixture is useful for keeping the header and IC socket in a fixed position while you work on it. A small bench or drill-press vise may be used as a holding fixture if you don't have a third-hand device.

A magnifying glass is almost mandatory when building MCMs. It will allow you to check frequently for unwanted solder bridges, poor joints and shorting leads on the header. To this end, a pencil soldering iron with a fine tip and low wattage rating (25-30 W) will help to minimize melting the IC-header plastic and the formation of unwanted solder bridges between the header terminals.

The first step in construction is to place all of the jumper wires on the header (as indicated by the pictorial drawings). Try to use light-gauge wire, preferably with insulation. The small wire from multiconductor telephone cable is excellent for this purpose. Bare wire may be used, provided there are no crossover jumpers on the header.

The general assembly procedure calls for installing the components of the MCM in layers. Some stacking will be necessary, depending upon the complexity of the circuit. After the jumper wires are in place, mount the resistors. If you can find some 1/8-W resistors, use them. This will minimize crowding on the header. I used ¼-W resistors for the examples shown photographically in this article, as I have no 1/8-W units in stock.

Next in the assembly comes the capacitors, followed by the transistors and, finally, the largest components. I add the RF chokes, toroidal coils and crystals last.

You may protect the tested, completed MCMs by developing a mold and encapsulating the components and header top side in casting resin. This will eliminate the possibility of replacing defective components later on, but it will keep dirt and moisture from entering the circuits. I use quick-setting epoxy cement for this purpose when I want to seal and anchor the components on some of my IC headers.

It is to your advantage to look for miniature components in the surplus catalogs and at flea markets. Large, old-style parts do not lend themselves well to MCM construction. Fortunately, the present electronics technology provides a substantial fallout of surplus mini components, and these are ideal for building MCMs.

A Marriage of MCMs

Four of the MCMs in this article may be used in concert to provide a simple D-C receiver. Fig 6 contains a block diagram of such an arrangement. The example suggests a circuit for 40-meter use. In this case, Z2 serves as a doubly balanced product detector. The output is at audio frequency

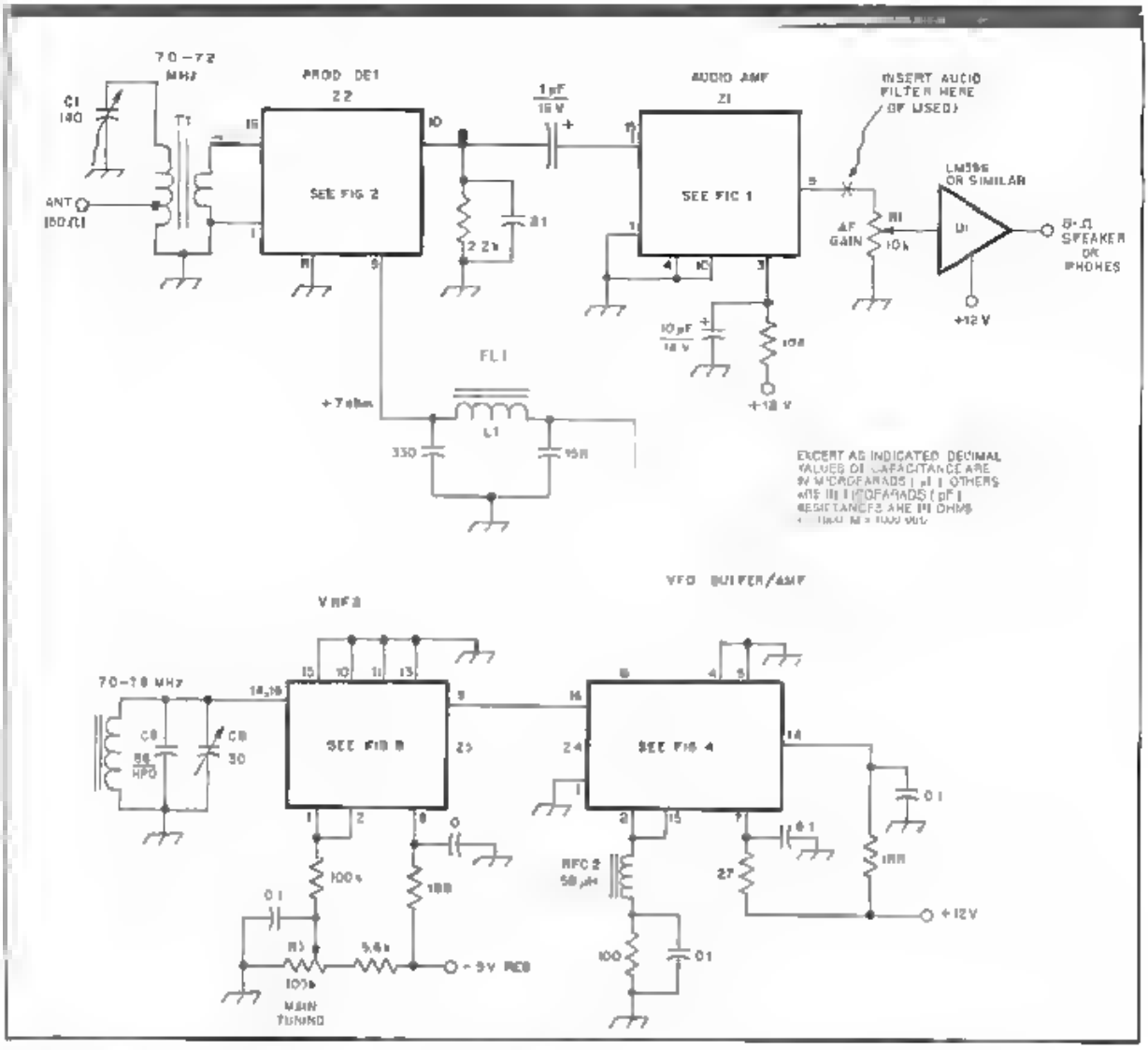


Fig 6—Suggested hookup for using four of the MCMs in this article in a 40-meter D-C receiver. See text for detailed information.

rather than at an IF, as would be the situation if Z2 were used as a mixer. Z1 operates as an audio preamplifier to drive an LM386 audio chip (or equivalent IC). Headphones may be connected to the output of Z1, but only strong signals will produce ample volume without U1 being added. D-C receivers require 80-100 dB of audio gain to permit weak-signal copy.

Z5 of Fig 6 is the VFO MCM, but when used in a D-C receiver it operates as a VBFO (variable beat frequency oscillator). C5 and C6 are used to set the operating range of the oscillator. R3 is the main tuning control for the receiver.

Z4 of Fig 6 amplifies the VBFO energy to an acceptable level for injection at pin 9 of Z2. FL1 provides an impedance match and offers some filtering of the Z4 output energy. A 10-20 dB RF preamplifier would be a welcome addition between the antenna and the input of Z2. This would greatly improve the receiver noise figure, and it should enhance weak-signal reception. A preamp may be built on an IC header to conform to the general format of the receiver.

CW reception will be greatly improved (better selectivity and reduced wide-band noise) by the addition of an LC passive or RC active audio filter at the point marked with an X at pin 9 of Z1. Suitable circuits may be found in the *ARRL Handbook for the Radio Amateur* and *Solid State Design* by the ARRL. Reception of SSB signals will be satisfactory without a filter, but the addition of a low-pass audio filter at X will improve the receiver selectivity for SSB reception.

An extremely compact receiver can be built by using the arrangement shown in Fig 6. With a few more MCMs of your design, it would be a simple matter to develop a small superhet receiver. You will need to build an IF-amplifier MCM and another one for the BFO.

Summary Comments

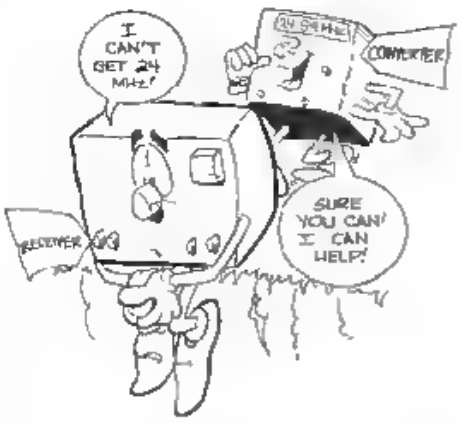
The intent of this article is to inspire you to try this method of miniaturization. I am told frequently that ham radio is supposed to be fun. Designing and building MCMs has been fun for me, and they offer some practical advantages over conventional PC board construction. You should be able to develop a bank of MCMs for various applications. This can hasten assembly time for many experimental circuits. MCMs are, of course, excellent units for use in a permanent circuit as well. Who knows, you may be the first ham in your area to create that elusive wrist radio of comic-book fame!

From April 1985 QST, p 42

A Converter for the 24-MHz WARC Band

Here's your chance to listen to a new band and enjoy an interesting construction project.

By Doug DeMaw,* W1FB



Perhaps you've wondered what is happening on the 24-MHz band but you can't listen to the frequency because your rig doesn't include WARC-band coverage. This converter is easy to assemble and get operating, and it's inexpensive.

Few RF circuits are laid out casually. Knowing how to approach the general design and assembly will be helpful in the years ahead when you build other RF projects.

General Design Objectives

First, ask "What do I want this converter to do?" Obviously, it needs to cover the band of interest — a foregone conclusion. But what of the other, sometimes subtle, considerations? Let's draft the criteria. The converter should:

1) exhibit an overall gain of unity, or slightly better. It should not create signal loss.

2) provide sufficient front-end selectivity to reject unwanted out-of-band signals.

3) be free of spurious responses and parasitic oscillations.

4) have a low noise figure (NF), permitting weak-signal reception.

5) have a dynamic range (ability to cope with strong in-band and out-of-band signals) that is reasonable to ideal.

At this point, you may be asking, "What does all of this really *mean*?" Well, let's examine the list, item by item.

It is possible to design a converter that exhibits a signal *loss*. This can degrade the signal-to-noise ratio (S/N) of the overall receiving system. A poor S/N ratio places the weak signals in the internal noise of the receiving system. This is similar, in effect, to having a normal signal become buried in atmospheric or man-made noise (QRN). Therefore, the converter must have ample gain and a low-enough NF to override the inherent noise of the receiver with which it is used. This does not mean that the converter must have an RF amplifier for all the amateur bands, but for 20 meters and higher it is wise to include one. Many converters for 40, 80 and 160 meters need only a mixer at the input stage, since atmospheric and man-made noise on those bands is usually greater than is the receiver noise.

Selectivity means that a tuned circuit or circuits with good Q (quality factor) should be used between the antenna and the first converter stage. This helps to discriminate against strong out-of-band signals. Some poor designs contain no tuned circuit ahead of the mixer, and that's an invitation to trouble!

To minimize spurious responses, you should ensure that no stage in a converter, other than the local oscillator (LO), is oscillating. The culprit in some home-constructed converters and receivers is the RF amplifier. Sometimes there is no outward indication of self-oscillation, and yet the unstable stage is generating a signal of its own. These random oscillations appear in the receiver output as unsteady or rough-sounding carriers, or "birdies." Under certain conditions, we may even find a mixer that is self-oscillating. Similarly, an oscillator may generate output on more than the desired frequency — especially if too much feedback is used. Other spurious responses can result from excessive harmonic output from the converter LO.

The transistor selected must be capable of providing an acceptable noise figure for our chosen operating frequency. This can be determined by looking at the manufacturer's data sheets for small-signal transistors that are earmarked for RF amplifier service. Let's be thankful that the NF requirements for 160 through 10 meters are not as stringent as they are at VHF and UHF! You can manage quite nicely with a maximum NF of, say, 5 dB in the HF spectrum. There are many transistors that meet this criterion. The f_T (upper-frequency limit versus gain) of a transistor must be correct, also. If not, the stage will not provide ample gain. I like to use a device that has an f_T of at least 10 times the operating frequency. For example, if I wanted to build a converter for 14 MHz, the RF amplifier transistor f_T would be 140 MHz or greater. Also, the noise figure is determined by the input-matching circuit and the specific biasing of the amplifier. So, you must take many things into account when designing an RF amplifier.

The dynamic range figure indicates the ability of the RF amplifier and mixer to handle large signals without generating IMD (intermodulation distortion), or going into gain compression (lowered gain). The system immunity to this is determined by the type of device used in the RF amplifier and mixer stages. Operating conditions based on dc voltages also play an important role in dynamic range. The approach you must take is anything but casual in this general area. There is a wealth of information on all five items on the list in the League's book, *Solid State Design for the Radio Amateur*.

What about Mixers?

There are so many pros and cons about mixer choice and operation that you could soon be wading in a sea of confusion if we discussed this subject in depth. The bottom line is to use a *strong* mixer; one that won't collapse when strong signals enter it. Diode-ring mixers (four diodes in a quad arrangement) are among the better choices, but they require more LO output power than is needed for a transistor or an IC mixer. Also, diodes operate as passive devices (no operating voltage is required), which results in a signal loss in the mixer. This is known as *conversion loss*. With a diode-ring mixer, the loss can be as great as 8 dB. The RF amplifier ahead of this mixer needs to have a gain of at least 10 dB to ensure a low noise figure. ICs such as the MC1596G and CA3028A offer good performance as

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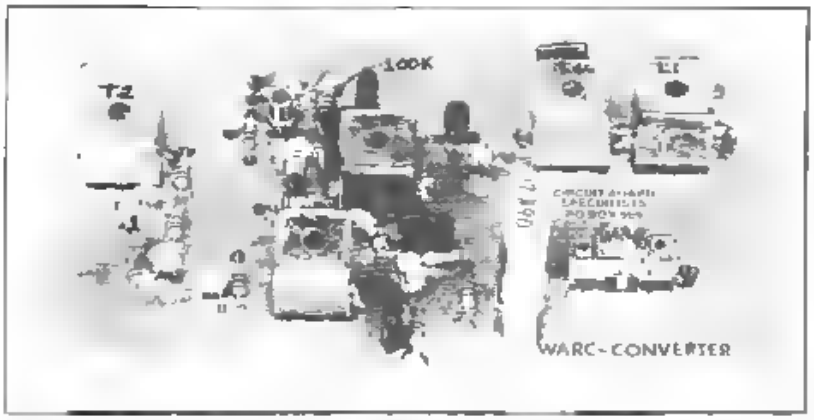


Fig. 1 — The assembled converter. Shielded Amidon Associates L43 and L57 transformer assemblies are used for the tuned circuits.

mixers. They do not require high LO power.

Another mixer requirement is that ample LO injection power or voltage be applied. Too little LO power to a mixer results in reduced gain and degraded dynamic range.

All of you won't grasp these fundamentals instantly. But you should have knowledge of the pertinent terms and a rough notion of what the terms relate to. I suggest further study in the various ARRL textbooks.

A Practical Converter You Can Build

Fig. 1 shows an assembled version of the circuit in Fig. 2. As shown, it is set up for operation in the 24-MHz band. PC boards and complete parts kits for this converter are available.[1]

In an effort to trade high performance for simplicity, I have chosen a design that uses only three transistors. Q1 is a grounded-gate (common gate) JFET RF amplifier. If the gate lead is kept very short when grounding it, the stage should be unconditionally stable. A good RF amplifier should not oscillate when the load is disconnected from J1. The stage gain is on the order of 10 12 dB. The same transistor, if used in a grounded-source hookup (input signal to the gate), can yield up to 20 dB of gain, but will be more difficult to tame.

A 40-meter trap (L1 and C1) help prevent 40-meter signals from riding through the converter. The 40-meter band is used as the tunable IF for this converter. T1 offers reasonable front-end selectivity. The source of Q1 is tapped near the ground end of the main transformer winding to provide an approximate 1:1 match between the 50-ohm antenna and 200-Ω source impedance of Q1. The source impedance of Q1 is determined by

$$Z = \frac{10^6}{g_m} \qquad \text{(Eq. 1)}$$

[1]Notes appear at end of article.

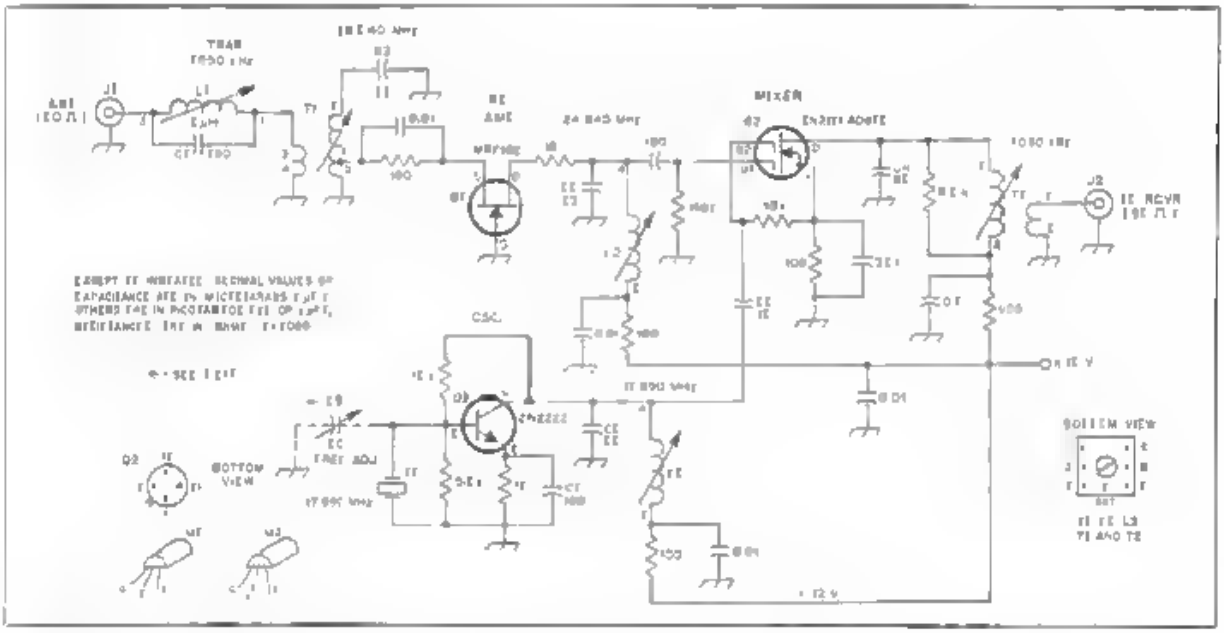


Fig. 2 — Schematic diagram of the 29 MHz receiving converter. Fixed-value capacitors are disc-ceramic, unless otherwise noted. Resistors are ¼ or ½ W carbon-composition.

C1-C7 incl. — Silver mica, polystyrene or NPO ceramic capacitors.
C8 — Optional 60 pF trimmer (see text).
J1, J2 — Phono jack, single-hole mount.
L1 — 5-μH (nom.) coil. Use 28 turns of no. 30 enam. wire on an Amidon Assoc. L43-6 transformer bobbin.
L2 — 0.75 μH (nom.) coil. Use 10 turns of no. 26 enam. wire on an L43-6 transformer bobbin.
L3 — 1.1 μH (nom.) coil. Use 12 turns of no. 26 enam. wire on an L43-6 transformer bobbin.
Q1 — Motorola MPF102 JFET or equiv. VHF type.
Q2 — RCA 40673 dual-gate MOSFET or TI 3N211.
Q3 — 2N2222 or 2N2222A NPN transistor or equiv.
T1 — 0.75-μH (nom.) secondary winding. Use nine turns of no. 26 enam. wire on an Amidon Assoc. L57-6 transformer bobbin. Tap at two turns above the ground end. Primary is a one-turn winding over the secondary.
T2 — 5-μH (nom.) primary. Use 22 turns of no. 26 enam. wire on an Amidon Assoc. L57-6 transformer bobbin. Secondary has three turns of no. 26 enam. wire. See inset drawing for pin locations for L1, L2, L3, T1 and T2.
Y1 — Fundamental crystal in HC-6/U holder, 20-pF load capacitance (International Crystal Mfg. Co. type GP or equiv.)

Amidon Associates, 12033 Otsego St., North Hollywood, CA 91607, tel. 213-760-4929.
International Crystal Mfg. Co., Inc., 18 North Lee, Oklahoma City, OK 73102, tel. 405-236-3781.

where g_m is the transconductance in siemens (formerly called mhos) of the transistor used.

Our mixer is a dual-gate MOSFET. This transistor is simple to use and offers average performance as a mixer in terms of dynamic range. A tuned transformer, T2, is used in the mixer output to provide an impedance match between the drain of Q2 and the 50-ohm receiver input. The 5.6-kΩ load resistor across T2 sets the impedance value of the drain circuit, and provides a relatively broad response across 100 kHz of the 40-meter IF. The resistor lowers the tuned-circuit Q.

Q3, a bipolar transistor, is the oscillator. Y1 is a fundamental crystal. The load capacitance for the crystal is approximately 20 pF. C8 is an optional trimmer capacitor that you may add to shift the crystal frequency in order to make the receiver dial match the received frequency. If the crystal is slightly off frequency, C8 may be needed. If the parallel arrangement for C8 is not satisfactory, move C8 so it is between the lower end of Y1 and ground, in series with Y1. The parallel arrangement will lower the Y1 frequency, while the series hookup will raise the Y1 frequency.

The RF injection voltage on the mixer (gate 2) should not exceed 6-V P-P (2.12-V RMS). A scope or an RF probe and voltmeter can be used to check the Q2 injection voltage. If it is too low, increase the value of C7. Similarly, decrease the value of C7 if the injection voltage if too high. A value of 4- to 6-V P-P is best for a dual-gate MOSFET mixer. Injection voltages greater than 6 can destroy the mixer transistor.

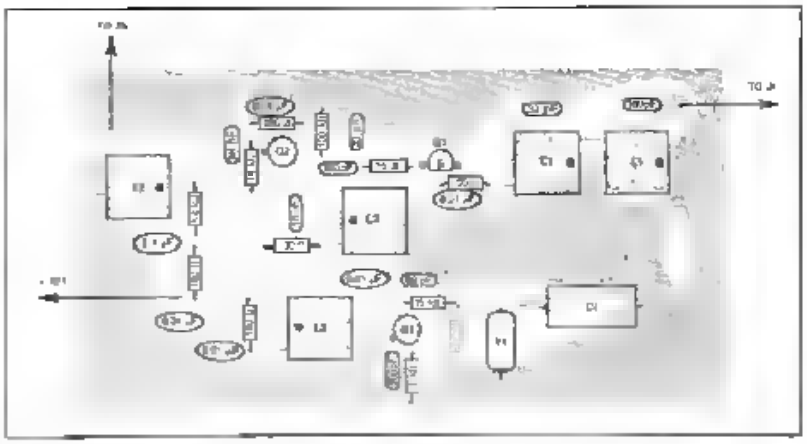


Fig. 3 — Parts-placement guide for the converter as seen from the component side of the board. The shaded area represents an X-ray view of the copper pattern.

Construction Notes

A parts-placement layout, seen from the component side of the board, is given in Fig. 3. A scale etching template is shown in Fig. 4.

If you decide to make your own circuit board, try to follow closely the pattern provided in this article. Double-sided board material is recommended for the circuit, but you may use single-sided material. Make certain that all solder joints are good ones. Component leads should be kept as short as possible.

Tune-up and Operation

The converter is capable of approximately 20 dB of gain when each tuned circuit is peaked for a single frequency on 40 meters. However, it is better to stagger-tune T1, L2 and T2 for 7.010, 7.050 and 7.075 MHz, respectively. Peak each circuit at the specified frequency. This will lower the effective converter gain somewhat while providing a more level gain response across the 100-kHz tuning range.[1] L3 is tuned for maximum output at 17.890 MHz. To ensure rapid starting of Q3, it may be necessary to tune L3 slightly off resonance to the high side of 17.890 MHz.

The 24-MHz amateur band extends from 24.890 to 24.990 MHz. Therefore, you will be listening to 24.890 MHz when your receiver is tuned to 7.0 MHz, and 24.990 MHz will be at 7.1 MHz on your receiver dial.

It is a good idea to enclose any converter in a shielded box. This prevents stray pickup of unwanted signals by the circuit board and various components. This is especially important in order to keep 40-meter signals out of the main station receiver during 24-MHz reception. Also, try to find a 40-meter signal that is leaking through the converter somewhere near 7050 kHz. Then adjust the trap, L1/C1, for minimum strength of the unwanted 40-meter signal.

Good luck, and see you sometime soon on 24.890 MHz!

Notes

Circuit Board Specialists, P. O. Box 969, Pueblo CO 81002, tel. 303-542-5083.

[1]Excessive converter gain can decrease the dynamic range of the receiver used as the tunable IF.

For updated supplier addresses see ARRL Parts Supplier List in Chapter 2.

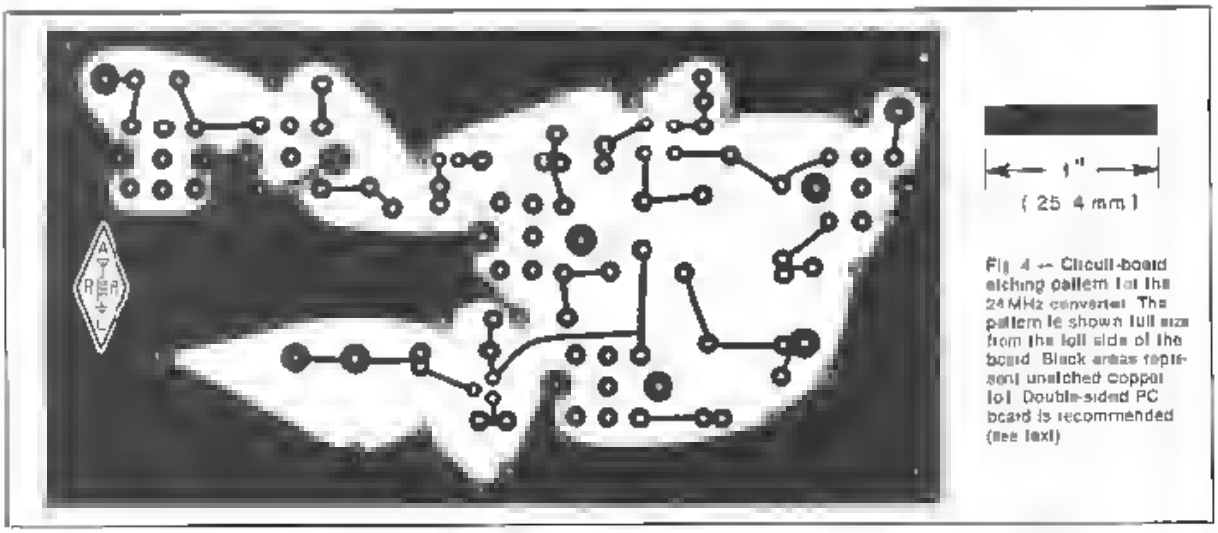


Fig. 4 — Circuit-board etching pattern for the 24-MHz converter. The pattern is shown full size from the foil side of the board. Black areas represent unetched copper foil. Double-sided PC board is recommended (see text).

From February 1989 QST, p 43:

ANOTHER ONE-MOSFET CONVERTER

Almost fifteen years ago, *QST* published a 10- and 15-meter converter that used a 40673 dual-gate MOSFET as mixer *and* crystal oscillator—a *converter* stage (see Fig 6).[1] Despite the article's report that the circuit oscillated reliably with ten different crystals, I recall having heard that some builders had trouble getting the circuit to work.

A variation on the single-MOSFET converter appears in the December 1987 issue of the Japanese magazine *CQ Ham Radio*. The Japanese configuration differs from McCoy's *QST* circuit in that a parallel tuned circuit (resonant at the crystal frequency) between the MOSFET drain and the output tuned circuit (resonant at the IF) is used to keep the drain impedance high at the crystal frequency. With sufficient separation between the crystal and intermediate frequencies, this drain *trap* should not unduly attenuate the converter's IF output. Fig 7 shows the circuit, along with component values for the working model I built in the ARRL Lab.

The values shown in Fig 7 have not been optimized, the tuned-circuit reactance values, in particular, were pulled out of thin air with the intent of constructing a working model quickly. The first crystal I found in my "junk box" was a 4-MHz microprocessor-clock unit; I chose the converter input and output frequencies (14 and 10 MHz, respectively) because they "work" with a 4-MHz LO.

Yes, it works. Dynamic range? I have no idea. Sensitivity? You've got me, although disconnecting my indoor antenna from the converter made most of the received background (not line) noise disappear (the "low tech" sensitivity test!). Image rejection? Not so hot; but this simple prototype has only one tuned circuit between the antenna and gate 1 of the MOSFET, after all.

How does the *CQ Ham Radio* circuit compare with McCoy's? Well, my Fig 7 prototype doesn't oscillate if the 4-MHz drain trap (L1 and the 82-pF capacitor) is shorted; shorting the drain trap of the *CQ Ham Radio* circuit approximates McCoy's hookup. (I suspect that the impedance of T2's resonant secondary is too low at 4 MHz to allow Q1 to "take off" without the drain trap. At some combinations of intermediate and LO frequencies, this may not be a problem. Crystal characteristics undoubtedly play a part.) The McCoy circuit uses positive bias on gate 2 of the MOSFET, and keeps gate 1 and the source at the same dc potential. The Japanese circuit returns both gates to ground; in conjunction with the voltage drop across the 270-Ω source resistor, this biases both gates negatively relative to the source. Even with positive bias applied to gate 2 of the MOSFET, however, my prototype does not oscillate with its drain trap shorted. Message: The drain trap is important! (*CQ Ham Radio* carried one version of this converter in which the gate-2-to-ground resistor was 10 kΩ instead of 100 kΩ; in that circuit, Y1 was a 41-MHz crystal, and a 5-pF feedback capacitor was connected from gate 2 of Q1 to ground.)

It pays to make L1, or its resonating capacitor, variable. In my prototype, the crystal oscillated on several frequencies at once and generated broadband hash unless the drain trap was tuned just so. But it was possible to find an L1 setting at which Q1 oscillated cleanly. In my opinion, this merely means more fun for the experimenter! (I also point out that we're perhaps being a bit unkind to the MOSFET in this circuit: Amplitude limiting—essential in any oscillator that does not destroy its active device[s]—obviously occurs *somewhere* in the circuit, but not by design! [Unlike the cathode-grid diode in a vacuum-tube oscillator, a MOSFET's gate-source insulator can't conduct without instantly destroying the device. Perhaps drain saturation is the amplitude limiter in this case.] McCoy reported that the highest RF voltage measured on gate 1 in his circuit

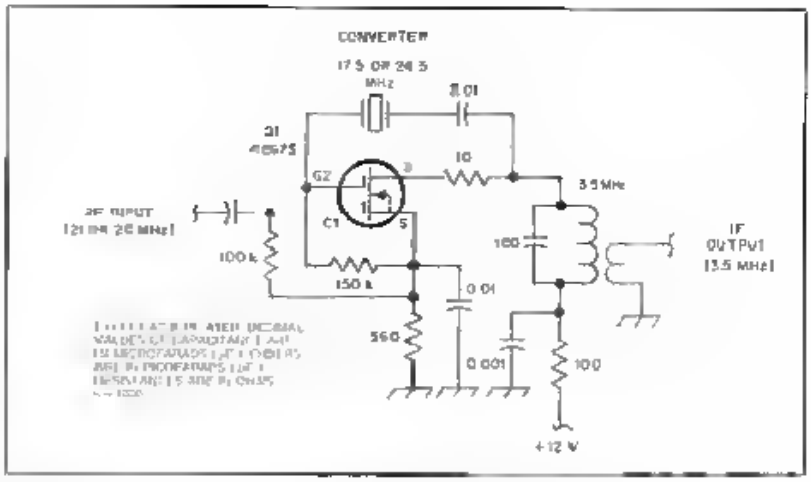


Fig 6—This one-transistor converter stage (RF amplifier not shown) appeared in March 1974 *QST*. Q1 acts as a Pierce oscillator and mixer. The 10-Ω drain resistor may have been included to suppress VHF parasitic oscillations in Q1.

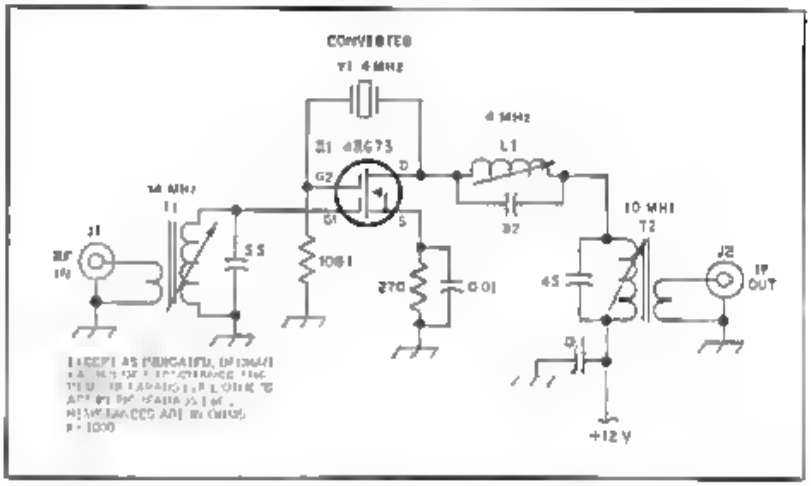


Fig 7—A one-MOSFET converter based on a *CQ Ham Radio* design. The principal difference between this circuit and that shown in Fig 6 is the drain trap (L1 and the 82-pF capacitor). Solenoidal slug-tuned inductors and transformers were used only because they were handy; their toroidal equivalents should work as well or better. See text.

J1, J2—coaxial RF connectors.

L1—14.8- to 31-μH slug-tuned coil (Miller 4407) set to approximately 19.3 μH. Reactance: approximately 485 Ω at 4 MHz.

T1—2.7- to 4.2-μH slug-tuned coil (Miller 4307) set to approximately 3.9 μH; primary: three turns of no. 26 enameled wire over cold end of secondary. Reactance of secondary approximately 344 Ω at 14 MHz.

T2—4.7- to 6.8-μH slug-tuned coil (Miller 4407) set to approximately 5.9 μH; secondary: four turns of enameled wire over cold end of primary. Reactance of secondary approximately 370 Ω at 10 MHz.

was 4—well within the ratings of the 40673. I did not measure the gate 1 voltage in the *CQ Ham Radio* circuit.)

Might this single-MOSFET converter work with overtone crystals? I dunno; *you* experiment, and tell us about it! How about configuring Q1 as an LC, instead of a crystal, oscillator? Great idea! Let me know your results.

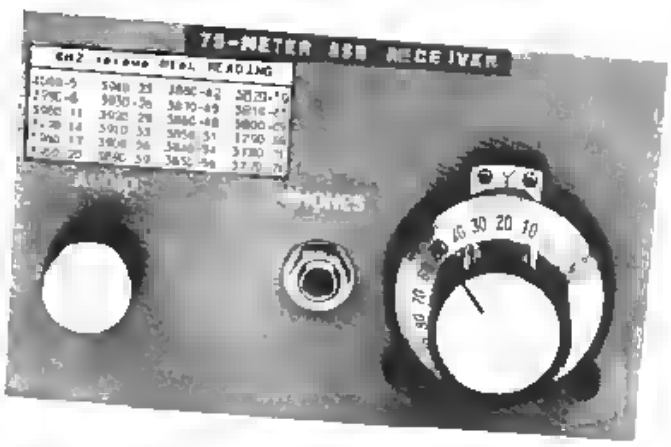
The circuit does what I wanted: It works—it "makes noise"—and it's interesting to fiddle with. Maybe you can find a good use for it. You might even have some fun along the way!

—*David Newkirk, AK1M, ARRL Staff*

[2]L. McCoy, "Improving Your Receiver Performance on 15 and 10 Meters," *QST*, Mar 1974, pp 26-27.

From May 1989 QST, p 25

A Four-Stage 75-Meter SSB Superhet



Getting "the most for the least" is a typical ham radio objective for those who build circuits. This simple SSB superhet receiver is the product of such an effort.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther, MI 49656

"Most everything you design be for CW operation?" I have been asked this question many times. Perhaps my preference for CW work influenced my thinking when I sat down to design a new piece of gear. The circuit in this article is my "apology" for overlooking the voice operators who like to build homemade receivers. I want to stress that this four-stage receiver does not belong in the high-performance class. However, it is sensitive and stable, and it provides good audio quality.

Design Rationale

One objective when starting this project was to learn how few components could be used to develop a receiver with acceptable performance. A great deal of editing and printing took place over a one-month period of breadboard-circuit testing. I wanted to have *some* rejection of the unwanted sideband, but I also wanted to minimize the cost of a crystal filter. A low overall noise figure was also a criterion.

Another goal was to have a surplus of audio gain for even the weakest of SSB signals. Finally, the power consumption for the receiver should be modest enough to permit battery operation during emergency or field use. All of these objectives have been met.

Circuit Highlights

The tuning range of the circuit in Fig 1 is from 3.7 to 4.0 MHz. The FL1 and oscillator constants can be changed to provide coverage of the 80-meter CW band, should you prefer that to the SSB segment of the band. Filter information is presented in *The ARRL Electronics Data Book*.[1] A slight increase in inductance is needed for L5 in order to cover 3.5 to 3.7 MHz.

Although Q1 could be made to work as both a mixer and oscillator, I chose to isolate the oscillator from the mixer. Harmonic currents also inject the mixer when both circuits share a common transistor substrate. This causes all manner of spurious responses, and oscillator pulling may also be a problem. The injection waveform from the gate of Q3 is very clean.

FL1 is a band-pass filter with circuit values taken from the W7ZOI tables in the *Data Book*. Although the values specified in Fig 1 are for 3.8 to 4.0 MHz, the attenuation at 3.7 MHz is minor with the filter peaked at 3.85 MHz. There is some insertion loss through FL1 (about 2 dB). An earlier version of this receiver had a single, high-Q tuned circuit at the mixer input. Receiver sensitivity was better with that arrangement, but it was a nuisance to retune the input circuit when changing frequency. With the single tuned circuit a 0.35-μV signal was 3 dB above the noise floor of the receiver. A 3-dB rise occurs at 0.55 μV with FL1 in place. I should mention also that the single tuned circuit allowed signals from the image side of the mixer (20 meters) to pass through the receiver. The band-pass filter corrected the fault.

Should you want to cover both the 75- and 20-meter bands you can build a 20-meter version of FL1 and band switch the two filters. As with the 75-meter-only version, an IF of 9.0 MHz (Y1) is required. With this arrangement the 20-meter band will tune backward from the 75-meter band, but upper- and lower-sideband reception will occur, as required, without changing the BFO frequency (Y2). This two-band scheme with a 5 MHz VFO is an old one!

In effect, the circuit in Fig 1 is a fixed-tuned direct-conversion receiver (Q2 and U1) with a tunable converter (Q1 and Q3) ahead of it. There are no IF amplifiers, and hence no AGC. Gain from an IF amplifier is not needed to ensure good performance. The overall receiver gain is approximately 75 dB. This is more than adequate for headphone reception.

Q2 serves as a crystal-controlled BFO and product detector. C14 is chosen to provide a BFO frequency that is roughly 1.3 kHz higher than the IF-filter crystal, Y1. (C12 and C13 lower the marked frequency of Y2.) A 50-pF trimmer can be used at W1/C14. You may want to eliminate C14 and order Y2 for a frequency that is 1.5 kHz higher than that of Y1. I found that I could shift a surplus 9.500-MHz HC-6/U crystal to 9.50013 MHz with C14 in place of W1 as shown. Changing C12 and C13 to 47 pF may help raise the Y2 frequency. I find that plated crystals in HC-6/U holders shift upward better than the small units in HC-18 holders. Crystals in FT-243 holders are not recommended for this application.

R3 is chosen to provide a relatively broad band-pass response for Y1. You may want to experiment with this value if you use crystals other than those listed in Fig 1. Filter ringing was a problem with a 100 kΩ value at R3. It appeared as a howl in the receiver output. C15 and C17 are used to prevent BFO energy from reaching U1. These capacitors also roll off the high frequency audio response to minimize the effect of high-pitched audio energy.

*Notes appear at end of article.

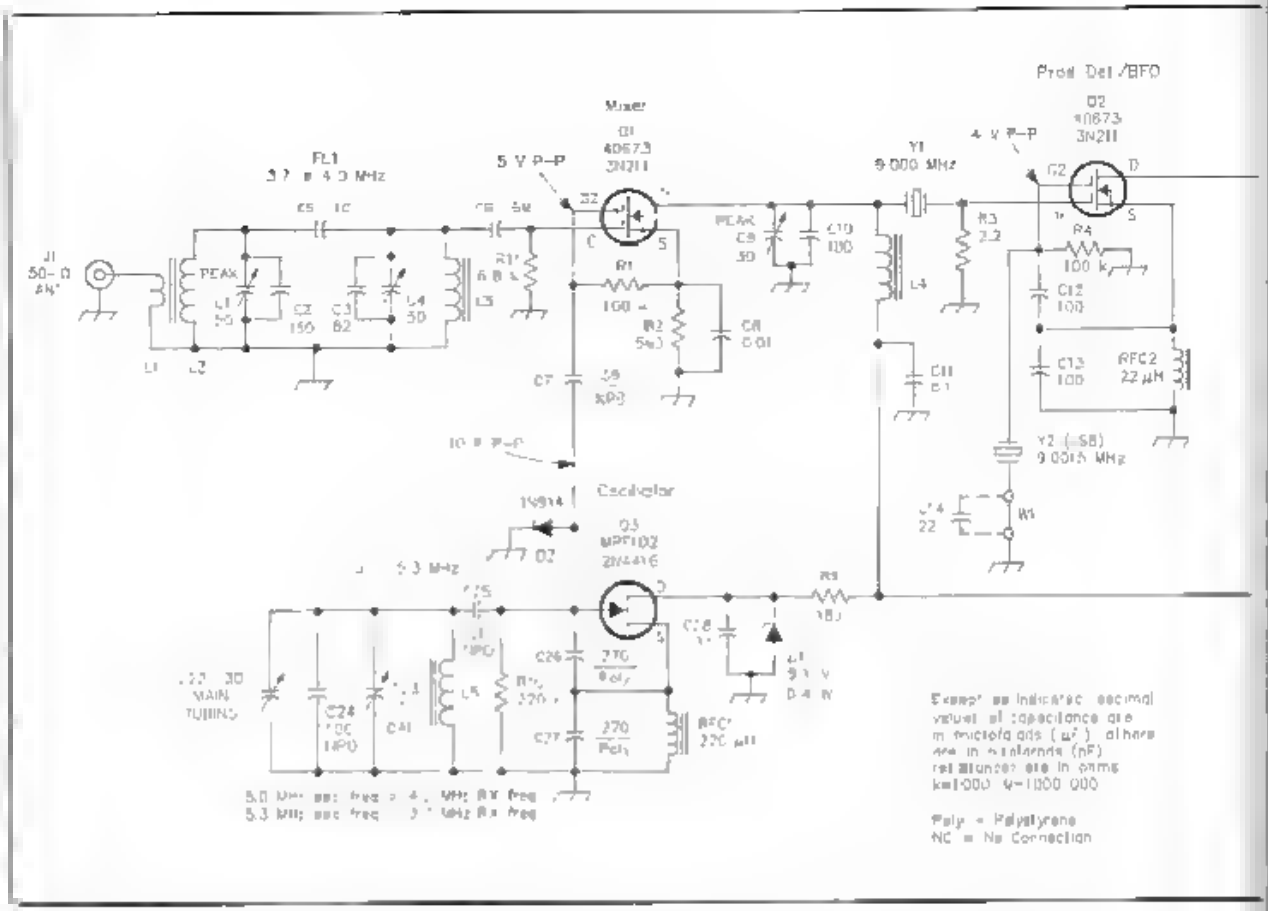


Fig 1—Schematic diagram of the simple SSB receiver. Fixed-value capacitors are disc ceramic unless otherwise noted. Polarized capacitors are tantalum or electrolytic. Fixed-value resistors are ¼-W carbon composition.

C1, C4, C9—Small plastic or ceramic trimmer, 50 or 60 pF. Small mica trimmers can be used also.

C2, C3, C10—Silver mica, polystyrene or NP0 disc.

C22—Miniature 30 pF air variable with shaft. Double-bearing type preferred. Should rotate smoothly.

C25—Ceramic trimmer capacitor, NP0 type preferred.

D1—9.1-V, 400-mW or 1-W Zener.

D2—Silicon high speed switching diode, type 1N914.

L1—Four turns of no. 25 enam. wire over L2 winding.

L2, L3—8.7-μH toroidal inductor, 44 turns of no. 28 enam. wire on an Amidon Assoc. T-50-6 toroid. Q_u = 170.

L4—2.6-μH toroidal inductor; 25 turns of no. 26 enam. wire on an Amidon Assoc. T-50-6 toroid.

L5—5.0-μH toroidal inductor, 33 turns of no. 22 enam. wire on an Amidon Assoc. T-68-6 toroid. Add two coatings of polystyrene "Q Dope" to winding for rigidity. Polyurethane varnish can be substituted.

Q1, Q2—Any dual-gate VHF MOSFET. RCA 40673 or 3N211 suitable.

R1—Audio-taper carbon-composition control.

RFC1, RFC2—Miniature ferrite-core RF choke (Mouser, see note 4).

Y1, Y2—5.0 MHz (Y1) and HC-6/U 9.0015-MHz (Y2) crystals (30 pF load capacitance). Available from JAN Crystals, 2341 Crystal Dr, PO Box 06017, Fort Myers, FL 33906, tel 800-237-3063. Catalog available.

The measured rejection of the unwanted (upper) sideband at 700 Hz (single tone) was 16 dB with a high-Q HC-6/U crystal at Y1. The closer the BFO frequency is to the IF, the worse the rejection. A two-crystal lattice filter can be substituted for Y1 if better rejection is desired. You may also want to consider a four-crystal ladder filter.[2]

Q3 operates as a Colpitts oscillator. C22 permits coverage from approximately 3.7 to 4.0 MHz. NP0 capacitors help to ensure acceptable long-term stability. NP0 units can be used at C26 and C27 to further improve the stability, although polystyrene capacitors are quite temperature stable.

Preventing Problems

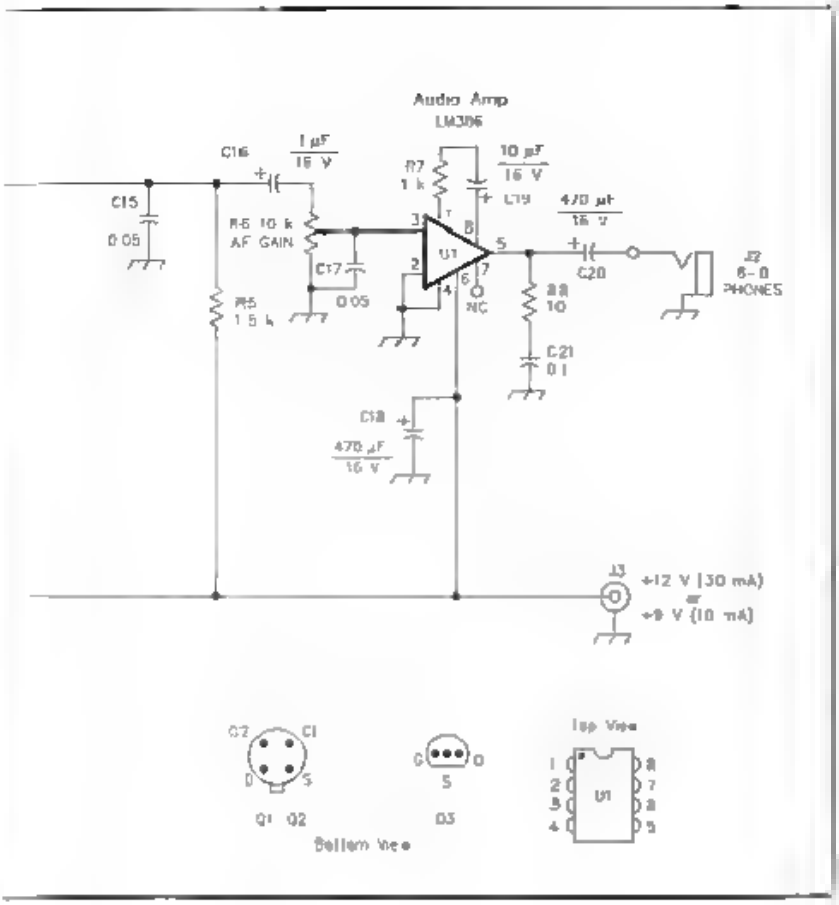
Owing to the high gain of U1, it is necessary to keep the leads going to the IC as short as practicable. C18 should be located as close to pin 6 as possible. R8 and C21 need to be close to pin 5 and C17 should be near pin 3. The gain of U1 can be increased by decreasing the value of R7, but instability lurks nearby when the chip gain is boosted!

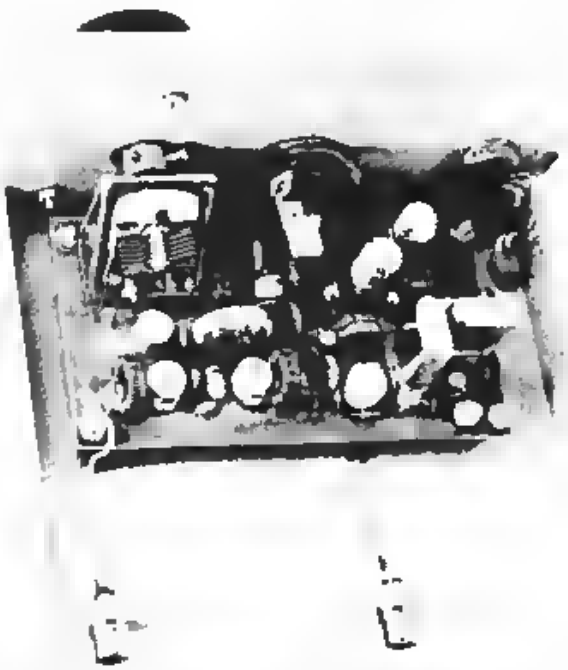
The value for RFC2 is critical. Too large an inductance value causes unwanted self-oscillation below 4 MHz. Use no more than 50 μH of inductance at RFC2.

Do not install Q1 and Q2 on the PC board until all of the other parts have been soldered in place. Dual-gate MOSFETs have fragile gate insulation, and static charges can perforate the insulation, thereby shorting the gates to the drain-source junction. Ground the tip of your solder pencil before soldering the FETs to the circuit board, and use minimum sustained heat.

Construction Comments

A PC-board etching pattern is provided in Fig 2. Boards for this project are avail-





Interior photo showing component layout. This circuit was designed with a minimum number of components, but optional modifications allow 20-meter coverage.

ARRL Lab Test Results

Tests of the model built by the author showed these results:

Minimum discernible signal (MDS) −99.0 dBm (decibels relative to a milliwatt) at 3800 kHz

Blocking dynamic range at 3800/3850 kHz 76.0 dB

Two-tone, third-order dynamic range at 3800/3850 kHz 59.0 dB

able from FAR Circuits.[3] A parts-placement guide is given in Fig 3. Single-sided PC board is used for this project.

Main-tuning capacitor C22 should be driven with a vernier mechanism to make tuning easy. An imported dial drive is suitable. The number scale can be used for frequency logging. Vernier drives are available by mail.[4] Surplus gear drives are available from dealers that sell WW II surplus.[5]

If C22 is not mounted securely to the receiver chassis or mainframe, cabinet flexing will cause mechanical frequency instability. Locate C22 as close to the Q3 circuit as possible.

Pads are available on the PC board for HC-6/U and HC-18/U crystals. The crystals can be soldered directly to the board, or you may install crystal sockets for Y1 and Y2. PC-board crystal pockets are available from International Crystal Manufacturing Co.[6]

Mount the receiver PC board by means of four metal spacers. This ensures that the ground foil of the board is well grounded to the mainframe. Proper grounding aids circuit stability.

Receiver Alignment

Use a frequency counter (or general-coverage receiver) coupled to RFC1 via a 50-pF capacitor to set 23 for the desired VFO range. The frequency range of the VFO is dependent on the crystal you select for Y1. In any event, it should have a 300-kHz range for coverage from 3.7 to 4.0 MHz. Adjust C22 for maximum capacitance and tweak C23 to obtain a 5.0-MHz reading on the counter.

Attach an antenna or signal generator to the input of FL1. Find a weak signal at approximately 3.85 MHz. Adjust C3 and C4 for maximum signal response. Repeat this step three or four times to overcome interaction between the resonators in FL1. Now, peak C9 for maximum signal level. There are no further adjustments, assuming that Y2 is on the proper frequency. You can check the Y2 frequency by sampling RF energy at the top end of RFC2 with a small-value capacitor.

Concluding Remarks

This receiver can serve as a foundation for further experimenting. For example,

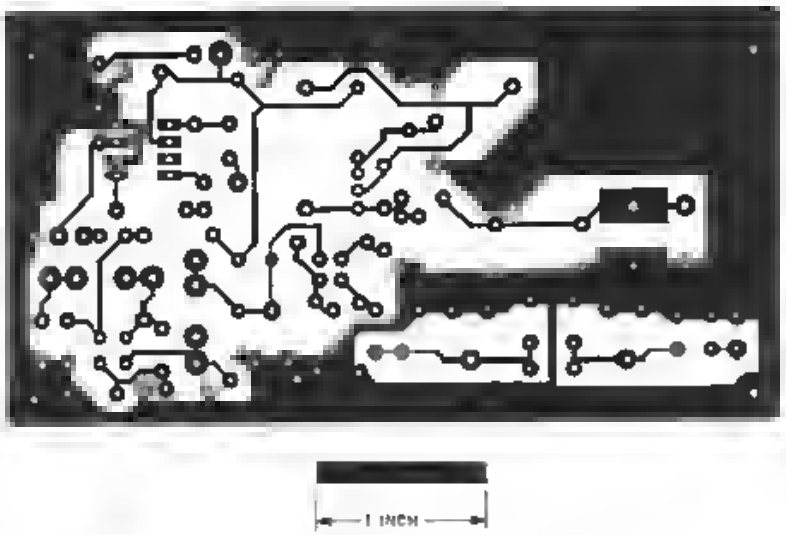


Fig 2—Circuit-board etching pattern for the receiver, shown full-size from the etched side of the board. Black areas represent unetched copper foil.

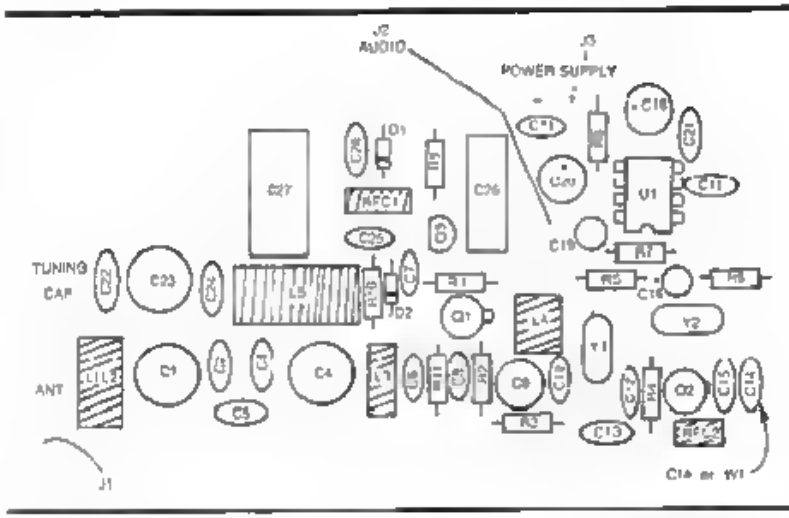


Fig 3—Parts-placement guide for the receiver. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. Component outlines are not necessarily representative of the shapes of the actual parts used.

the IF filter can be improved, as discussed earlier. An RC active audio filter can also be added to improve the overall receiver selectivity.

Although this receiver will drive an 8-ohm speaker rather well for loud signals, it falls short of the mark on weaker signals. This can be corrected by inserting a one-stage audio amplifier between Q2 and U1. A 2N3904 or 2N2222 can provide the extra gain needed for speaker operation. If this is done, add a 100-Ω resistor and 22-µF bypass capacitor to the supply lead that feeds Q2. This will decouple the audio circuits and prevent motorboating.

An S meter can be added by sampling the audio signal at the drain of Q2. Amplify the sampled audio with a 741 op amp, then rectify it with a 1N914 diode and filter it. A microammeter can be driven with the rectified audio to produce meter readings.

A class-A broadband RF amplifier can be added between FL1 and Q1 to enhance the receiver sensitivity. Circuits for these optional changes are given in Solid State *Design for the Radio Amateur* (ARRL).[7]

It should be a simple matter to modify this receiver for operating on other amateur bands. Only the VFO and FL1 need to be changed.

Notes

[1]Deleted

[2]W. Hayward, "Designing and Building Simple Crystal Filters," QST, July 1987, p 24.

[3]$6 (including shipping to US addresses) from FAR Circuits, 18N640 Field Ct., Dundee, IL 60118, tel 312-428-2431, evenings.

[4]Mouser Electronics, 2401 Hwy 287 N, Mansfield, TX 76063, tel 817-483-4422. Catalog available.

[5]Fair Radio Sales Co, 1016 E Eureka St, PO Box 1105, Lima, OH 45802, tel 419-223-2196. Catalog available.

[6]International Crystal Manufacturing Co, Inc, PO Box 26330, 701 W Sheridan, Oklahoma City, OK 73126-0330, tel 405-236-3741.

[7]Deleted

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

From 1990 ARRL Handbook Ch 30, p 43

A VXO-Controlled CW Transmitter for 3.5 to 21 MHz

The 6-W CW transmitter shown in Figs 47 to 50 can be built in a few evenings and will provide hours of on-the-air enjoyment. It features a variable-crystal oscillator (VXO) to generate a highly stable, adjustable-frequency signal. With the circuit shown here, frequency spans of 1 kHz or more can be realized. See Table 1. Only a few crystals are necessary for coverage of the popular CW frequencies. This single-band transmitter may be built for any one band from 80 through 15 meters. Since most crystals for frequencies above 25 MHz are overtone types, and this transmitter requires fundamental-type crystals, there is no provision for 10-meter operation.

Circuit Description

The schematic diagram of the transmitter is shown in Fig 48. Q1 and associated components form a Colpitts variable frequency crystal oscillator. C1 is used to adjust the frequency of the oscillator, and C2 is used to limit the span of the oscillator. If no limit is provided, the oscillator can operate "on its own" and no longer be under the control of the crystal. This is undesirable. On the 30-, 40- and 80-meter bands, C2 is not necessary and is omitted from the circuit. Supply voltage is fed to the oscillator only during transmit and spot periods. This prevents the oscillator from interfering with received stations operating on the same frequency.

Output energy from the oscillator is routed to Q2, a grounded-base amplifier. This stage provides some gain, but more important, it offers a high degree of isolation between the oscillator and the driver stage. Oscillator pulling and chirp are virtually nonexistent.

The driver stage uses a broadband amplifier that operates class A. This stage is keyed by grounding the base and emitter resistors. C10 is used to shape the keying waveform. Although the keying is rather hard, there is no evidence of clicks.

Two MRF476 transistors are used in parallel for the power amplifier. These transistors were designed for the Citizens Band service and work nicely at HF frequencies. Each transistor is rated for 3 W output. The original transmitter design used MRF472 output transistors, but Motorola no longer manufactures these devices. They are still available from many surplus outlets, however. L2 is used as a dc ground for the bases, making the transistors operate class C.

The low output impedance at the collectors of the output transistors is stepped up to 50 ohms by broadband transformer T3. A five-element Chebyshev low-pass filter is used to assure a clean output signal. This transmitter exceeds current FCC spectral purity specifications (see Fig. 49). D2 is used to clamp the collector voltage waveform to protect the output transistors if the transmitter is operated into an open circuit or high-SWR antenna system. The transmitter is designed to operate into a load that is close to 50 ohms resistive. S1 is used as the transmit/receive switch. One section transfers the antenna to an accompanying receiver or to the output of the transmitter. Another section is used to activate the VXO during transmit and the third section is provided for receiver muting purposes.

D3 and the associated components form an RF output driver for M1. This circuitry

Table 1

Component Values for the VXO-Controlled, 6-Watt Transmitter

Band	C1 (pF)	C2 (pF)	C3 C4 (pF)	C5 (pF)	C6, C7, C8 (pF)	L1	L3, L5	L4	VXO Range
80 M	365	*	220	100	820	47 Turns T50-2	25 Turns T50-2	32 Turns T50-2	3- 5 kHz
40 M	365	*	100	100	470	36 Turns T50-2	17 Turns T50-2	21 Turns T50-2	5- 8 kHz
30 M	150	*	68	50	330	27 Turns T50-2	14 Turns T50-2	18 Turns T50-2	8-10 kHz
20 M	50	10	50	50	240	30 Turns T50-6	14 Turns T50-6	17 Turns T50-6	10-12 kHz
15 M	50	10	33	33	150	23 Turns T50-6	11 Turns T50-6	14 Turns T50-6	12-14 kHz

*Not used

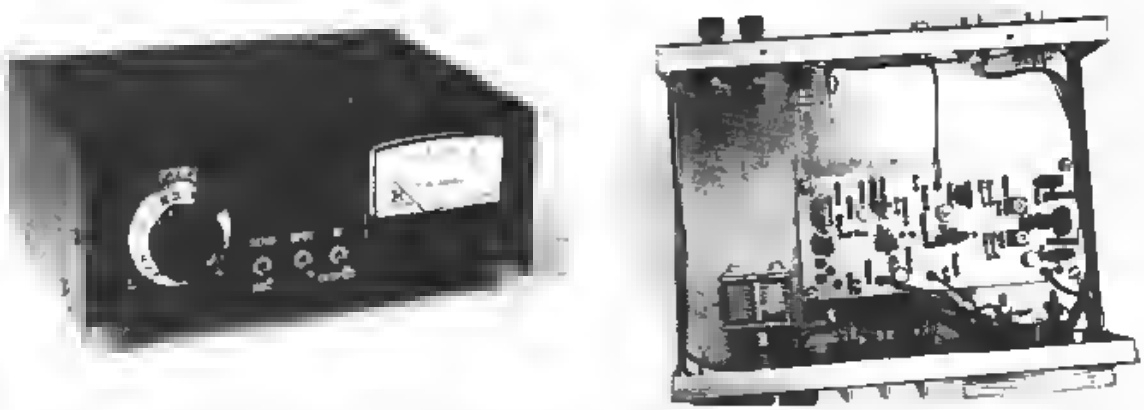


Fig 47 — The completed 6-W VXO controlled transmitter is housed in a small aluminum enclosure. RG-174 is used for all interconnections carrying RF. This version uses the older MRF472 output transistors, which are mounted flat on the board.

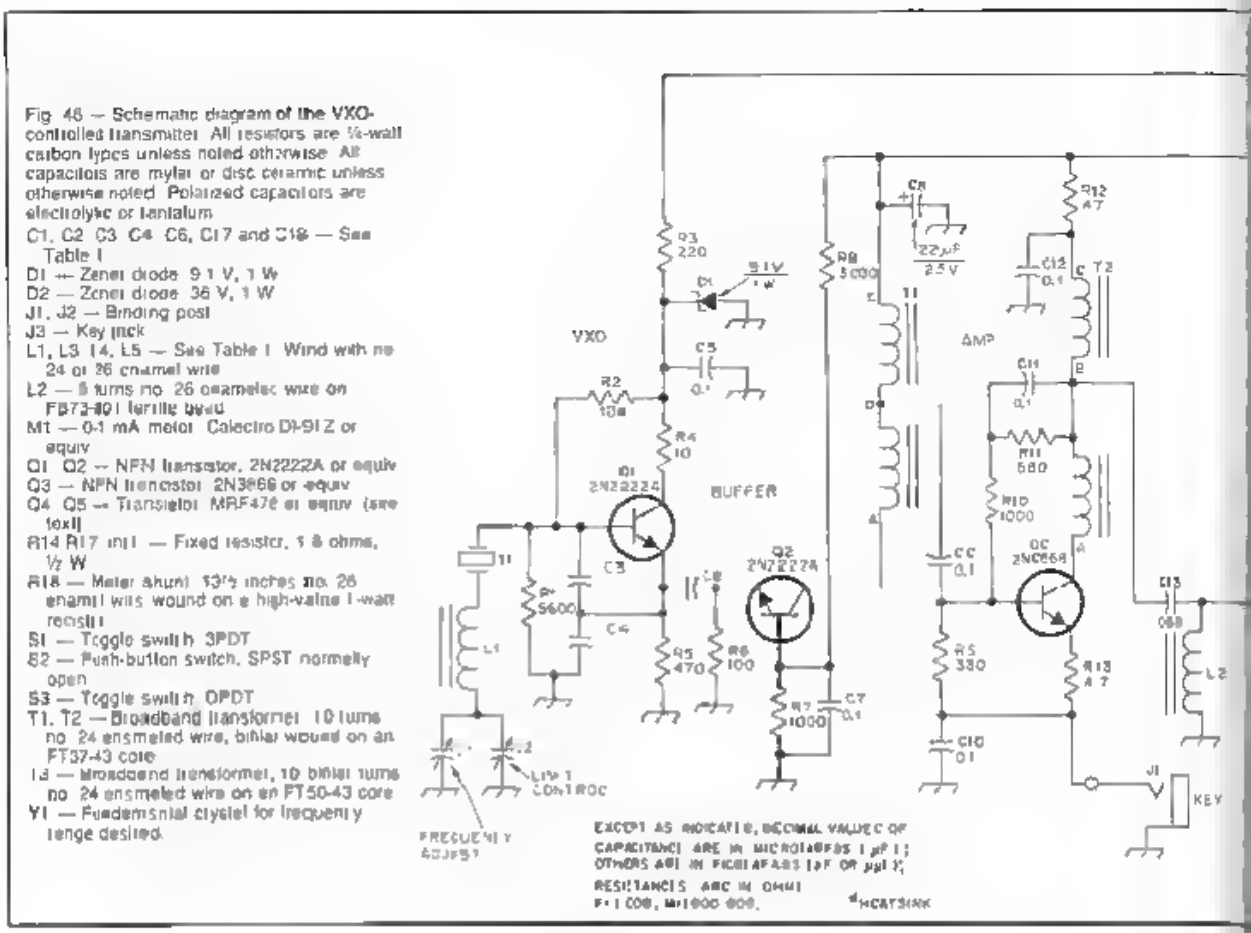


Fig 48 — Schematic diagram of the VXO-controlled transmitter. All resistors are ¼-watt carbon types unless noted otherwise. All capacitors are mylar or disc ceramic unless otherwise noted. Polarized capacitors are electrolytic or tantalum.

C1, C2, C3, C4, C6, C17 and C18 — See Table 1.
D1 — Zener diode, 9.1 V, 1 W.
D2 — Zener diode, 36 V, 1 W.
J1, J2 — Binding post.
J3 — Key jack.
L1, L3, L4, L5 — See Table 1. Wind with no. 24 or 26 enamel wire.
L2 — 8 turns no. 26 enamelec wire on FB73-801 ferrite bead.
M1 — 0-1 mA meter. Calectro D1-912 or equiv.
Q1, Q2 — NPN transistor, 2N2222A or equiv.
Q3 — NPN transistor, 2N3866 or equiv.
Q4, Q5 — Transistor, MRF472 or equiv. (see text).
R14, R17 (not) — Fixed resistor, 1.8 ohms, ½ W.
R18 — Meter shunt, 13½ inches no. 26 enamel wire wound on a high-value 1-watt resistor.
S1 — Toggle switch, 3PDT.
S2 — Push-button switch, SPST normally open.
S3 — Toggle switch, DPDT.
T1, T2 — Broadband transformer, 10 turns no. 24 enameled wire, bifilar wound on an FT37-43 core.
T3 — Broadband transformer, 10 bifilar turns no. 24 enameled wire on an FT50-43 core.
Y1 — Fundamental crystal for frequency range desired.

is optional as there are no power-output tuning adjustments. M1 is also used to monitor transmitter current consumption.

Construction

The majority of the circuit components are mounted on a double-sided PC board. One side of the board is etched with the circuit pattern, and the other side is left unetched as a ground plane. A small amount of copper is removed from around each hole on the ground-plane side of the board to prevent leads from shorting to it. A test transmitter was built in the ARRL lab using single-sided board and the transmitter seemed to function normally with no instability. No long-term testing was performed, however. A parts layout guide and photo of the finished board appear in Fig. 50.

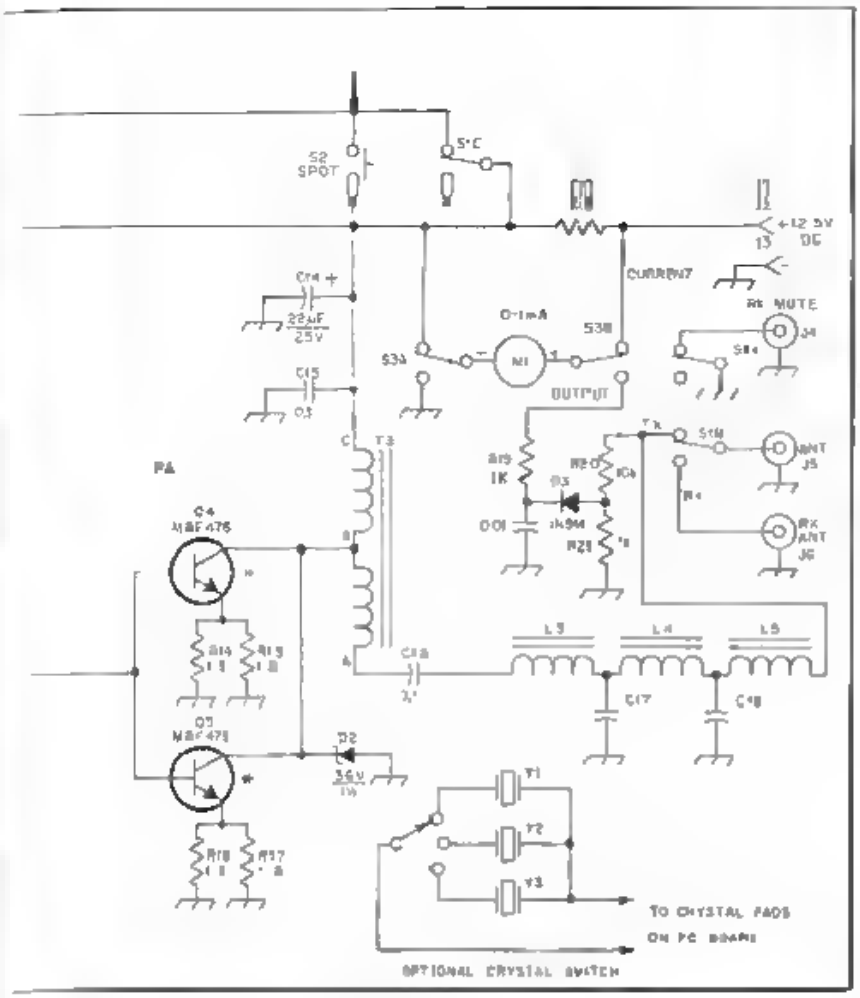
Affixed to the front panel are the transmit/receive switch, spot switch and the tuning capacitor. The rear apron supports the antenna and mute jacks, key jack and binding posts.

A homemade cabinet measuring 3 × 6 × 8½ inches was used in the construction of this transmitter. The builder may elect to build a cabinet from sheet aluminum or circuit-board material. The layout is not critical except that the lead from the circuit board to C1 should be kept as short as possible — an inch or two is fine.

A bent aluminum heat sink was attached to the output transistors. Commercial TO-220 heat sinks could also be used. If MRF472 transistors are available, they can be mounted flat on the circuit board and screws passed through the center of the transistors to hold them down. The ground plane will act as a heat sink sufficient for short key-down periods.

The only adjustment needed is that of setting the VXO limit capacitor (C2), and even this adjustment is not needed for the 80-, 40- and 30-meter transmitters. This adjustment can be done with the aid of a receiver. With a fundamental crystal in the circuit, adjust C2 for a maximum frequency spread that approximates the value shown in Table 1. If too much frequency spread is available, increase the amount of capacitance. Make a final check with the receiver by listening to the keyed signal from the transmitter. It should be steady and chirp free.

To provide wider frequency coverage, several crystals may be used. A crystal socket may be mounted on the front panel, or several sockets can be mounted on a separate circuit board and a simple rotary switch used to connect the desired crystal into the circuit. This option is shown in Fig. 48. Any number of crystals may be used, depending on the number of positions on the rotary switch. With crystals spaced 10-kHz apart, the circuit can provide continuous coverage of 50 60 kHz of the 20-meter band.



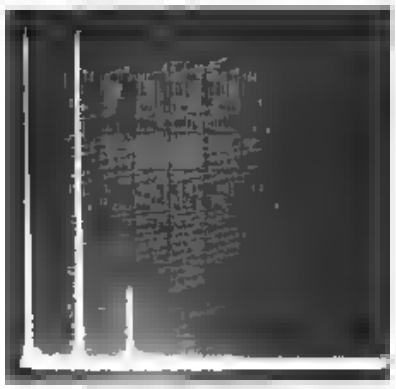
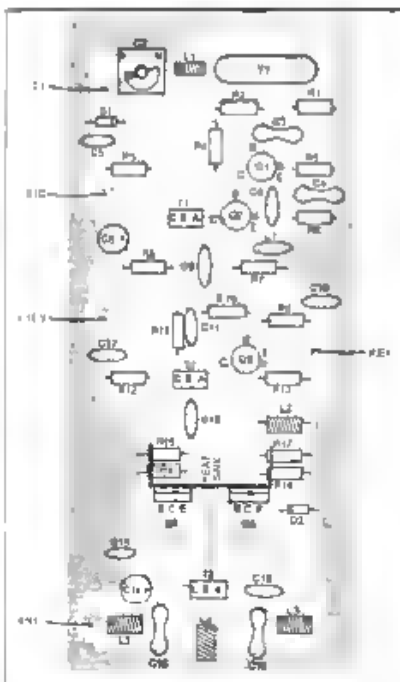


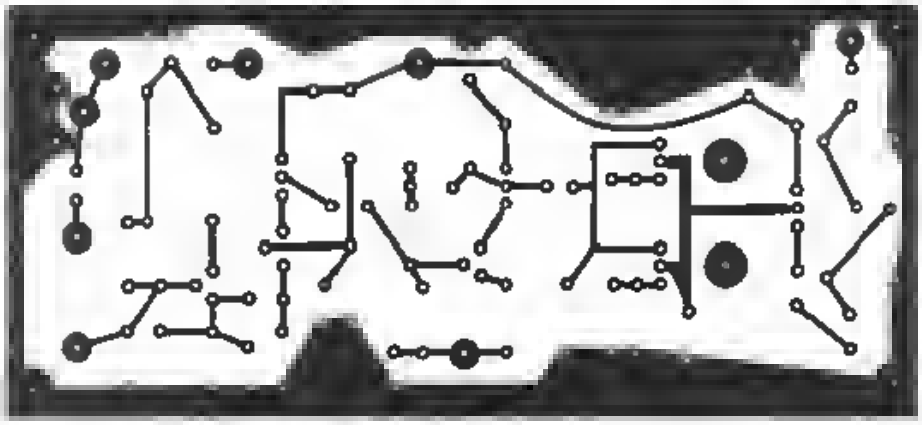
Fig 49 — Spectral display of the VXO-controlled transmitter. Here the transmitter is operated in the 20-meter band. The second harmonic is down 56 dB relative to the fundamental output. Similar presentations were obtained on each of the other bands. This transmitter complies with the current FCC specifications regarding spectral purity.



(A)

(B)

Fig 50 — The component-placement diagram for the 6-W transmitter PC board is shown at A. The component side of the board is shown with an X-ray view of the circuit foil. A full-size etching pattern appears at the back of this book. At B is a photo of the transmitter circuit board built using MRF476 transistors. The devices are mounted upright on the board with a heat sink attached to the metal tabs on the transistors.



Etching pattern for VXO-controlled CW transmitter. Shown full size from the foil side of the board. Black areas represent unetched copper.

From November 1989 QST, p 37:

18-MHz COMPONENT VALUES FOR THE *HANDBOOK* VXO CW TRANSMITTER

Yes, the *1989 ARRL Handbook*'s 6-watt, VXO-controlled CW transmitter works well at 18 MHz. Here are component values necessary for using the rig on this band; the component designators listed are those shown in Fig 48 of the *Handbook* write up:

C1—VXO tuning capacitor; 50 pF.

C2—Limits the VXO tuning range to ensure that the crystal and not L1 and C1, controls the oscillator frequency. I omitted this capacitor in the version I tested; if you try this and your crystal loses control, use 10 pF.

C3, C4—VXO feedback capacitors, 39 pF, silver mica or NP0 ceramic.

C6—Interstage coupling capacitor, 39 pF, silver mica or NP0 ceramic.

C17, C18—Output filter capacitors, 180 pF, silver mica (10 pF in parallel with 180 pF).

L1—VXO inductor; 28 turns of no. 26 enameled wire on a T-37-6 toroidal, powdered-iron core (measured inductance, 2.5 µH). Space the turns on this coil, and those on L3-L5, to allow a 30° gap between the beginning and end of the each winding.

L3, L5—Output filter inductor; 16 turns of no. 24 enameled wire on a T-37-6 core (measured inductance, 0.85 µH).

L4—Output filter inductor; 20 turns of no. 24 enameled wire on a T-37-6 core (measured inductance, 1.28 µH).

Y1—Parallel-resonant fundamental crystal, 20- or 32-pF load capacitance. An 18.07-MHz crystal borrowed from Zack Lau's QRP Tuna-Bander (see pp 25-30 of October 1989 *QST*) provided a VXO swing of 10.8 kHz with 39 pF at C3 and C4.

Powered with a 12.0-V dc supply, my version of the VXO transmitter draws 1.26 A dc while producing 6.2 W output at 18.09 MHz. Fig 4 shows the transmitter's output spectrum under these conditions.—*David Newkirk, AK7M, ARRL Staff*

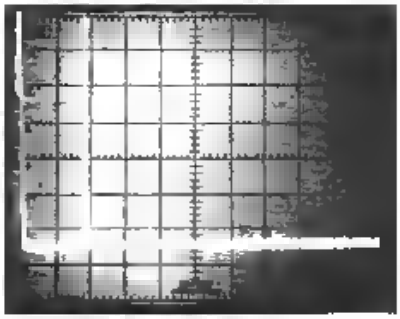
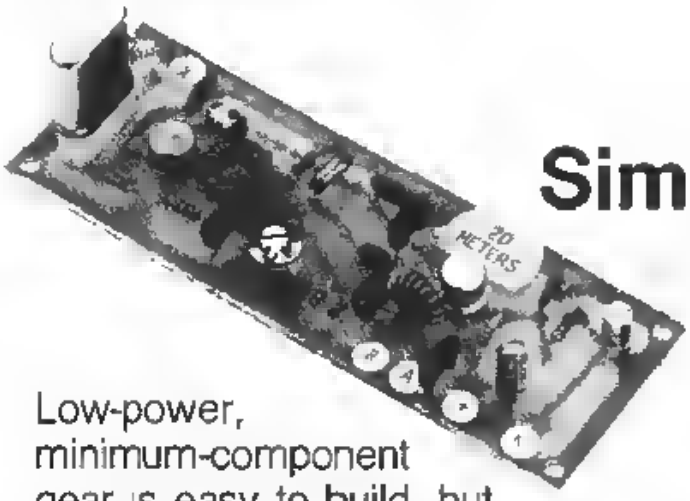


Fig 4—Spectral display of the *ARRL Handbook* 6-W VXO transmitter operating at 18.09 MHz. Each horizontal division represents 10 MHz; each vertical division represents 10 dB. The spike at far left (the spectrum analyzer's first local-oscillator signal) serves as a convenient "0 MHz" reference. This spectrogram was taken with the VXO transmitter producing 6.2 W of RF energy. All harmonics and spurious emissions are at least 57 dB below peak fundamental output. Modified for 18 MHz as described in the text, the 6-W VXO transmitter complies with current FCC specifications for spectral purity.

From January 1985 QST, p 22:



Simple QRP Gear Versus Good Performance

Low-power, minimum-component gear is easy to build, but performance is often below that which can be obtained with careful design. Let's consider some design pitfalls and the practical project offered here.

By Doug DeMaw,* W1FB

I'm sure you've heard a number of chirpy or buzzy signals from homemade QRP transmitters. In fact, you may have unwittingly turned loose a "super blooper" on the airways yourself! I'd be shading the truth if I did not confess to being guilty of a similar misdemeanor once or twice in my amateur career. It seems that the simpler the transmitter the more prone it is to chirp and related ills. Once we understand the causes of inferior performance, preventive steps can be taken in the design period to avoid generating "lid" style signals.[1]

Although we are addressing simple transmitters in this discussion, the general approach to design can be applied to most oscillators and VFOs that are part of a larger circuit, such as a multistage high-power transmitter. Since the oscillator is the heart of a transmitter or receiver, it deserves special care in the design and construction stages.

Profile of an Oscillator

There have been many discussions of what oscillators are and how they function. Certainly, *QST* and the ARRL *Handbook* have carried a wealth of data on this subject over past decades. But for the purpose of quick review, let's look at Fig. 1. The circuit example at A shows a Pierce oscillator that has an untuned output circuit. Feedback to make the circuit oscillate is between the base and collector of Q1.

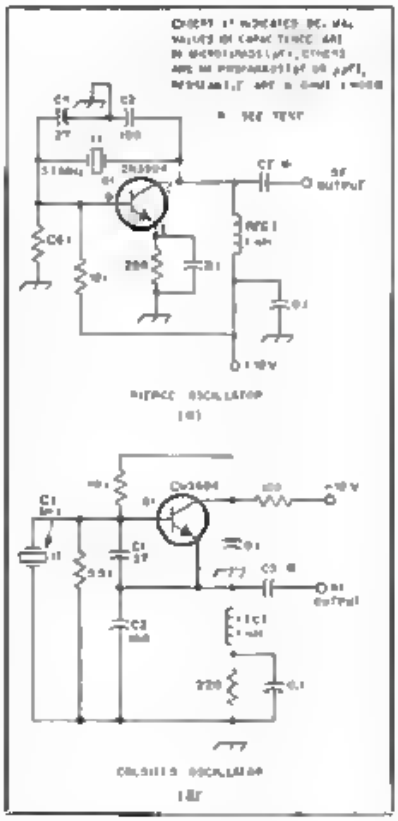


Fig 1 — Example of a Pierce solid-state oscillator (A). C1 and C2 control the feedback in the circuits at A and B. A Colpitts oscillator is found at B of this drawing.

This positive feedback is provided by C1 and C2. The capacitance ratio is adjusted to ensure fast starting of the oscillator when it is turned on or keyed. Remember that an oscillator is actually an amplifier. Part of the power output is routed back to the input circuit to cause oscillation. Therefore, C1 and C2 are chosen to ensure just enough feedback energy to provide reliable oscillation. Typically, a ratio of 4:1 in the feedback capacitors is a good starting value during initial design. In other words, we should attempt to use one fourth the output power as feedback energy.

We need to remember, also, that this feedback power is taken from the output power available for delivery to the load. It is for this reason that an oscillator is not as efficient (dc input power versus RF output power) as a straight RF amplifier. In the interest of best efficiency, we should use no more feedback power than is required for reliable circuit oscillation. There are other reasons why too much feedback is undesirable. It can cause a chirpy output signal and oscillation at frequencies other than the desired one, and may harm the crystal from the effects of excessive current. The high current can cause the crystal to heat up or even fracture. This danger is more pronounced as the operating frequency is increased (raised) because the higher the crystal frequency the thinner the quartz element in a crystal. An overheated crystal will drift in frequency, just as a VFO will with changes in component temperature. Some amateurs attempt to generate substantial power by using a power

[1]Notes appear at end of article.
*ARRL Contributing Editor, P.O. Box 250, Luther, MI 49656

oscillator in a one-stage transmitter. The results are often dismal, owing to excessive crystal current and drift.

Further examination of Fig. 1A shows that the emitter of Q1 is at RF ground by virtue of the 0.1-μF emitter bypass capacitor. Fundamental-frequency oscillators (output frequency the same as the crystal frequency) have only two terminals that are "hot" at RF. Attempts to have all three terminals "alive" with RF energy will result in unstable performance, or no oscillation at all. The exception is when the oscillator serves also as a frequency multiplier, as in Fig. 2.

RFC1 of Fig. 1A is selected to be self resonant well below the desired oscillator frequency. In our circuit, we find a value of 1 mH. We can assume approximately 10 pF of stray circuit capacitance at the output of Q1. Therefore, the self-resonant frequency of RFC1 is on the order of 1.6 MHz — well below 3.5 MHz, the oscillator frequency. Owing to the nature of this circuit, the output circuit of our oscillator is broadband. It would be suitable also for use above 80 meters if the values of the feedback capacitors were changed.

The amount of feedback energy needed is dependent mainly upon how active the particular crystal is, plus the beta of the transistor we happen to connect to our circuit. The beta (gain) of a transistor varies considerably between one transistor and another from the same manufacturer's production run. This is why the data sheets list the beta with a "typical" value. In reality, it can be above or below that value by a fair margin. Because of unknown crystal characteristics and the so-called "beta spread," we need to adjust the feedback for each circuit we build — if optimum performance is desired.

Colpitts Oscillator

Thus far we have considered only the Pierce oscillator. There are countless other types of crystal-oscillator circuits, and each is named after the person who developed it. The Pierce and Colpitts circuits seem to be the most common in amateur circuits, and that is why I have selected them for this discussion. The basic form of Colpitts oscillator is shown in Fig. 1B. In this example, we find that the collector is "cold" in terms of RF energy by means of the 0.1-μF bypass capacitor. The base and emitter terminals of Q1 are hot with regard to RF energy. C1 and C2 comprise the feedback divider. They can be adjusted in value to provide the required amount of feedback power.

RF output is taken from the oscillator emitter circuit. The output voltage is quite low compared to that of the Pierce circuit of Fig. 1A because of the low impedance of the Colpitts oscillator output tap point. Both oscillators require the smallest practical value of output coupling capacitor (C3) to minimize loading of the oscillator, which can cause chirp (when keyed) or no oscillation at all. A typical value for C3 at 80 meters is 50-100 pF. Smaller values are recommended for 7 MHz and higher. The larger the value of C1 in Fig. 1B, the greater the feedback amount. In some circuits, we may eliminate C1 completely. This is because there may be sufficient capacitance within the transistor to serve as C1. This will depend on the transistor type we select for Q1.

We can conclude from the previous discussion that there are many variables that dictate how we select component values for a given crystal oscillator circuit. Experimentation has long been the motto of amateurs, so this requirement should be a matter of course for most of us who manipulate a soldering iron in the small hours of the morning! The variables that apply to crystal oscillators do, of course, relate to VFO (variable-frequency oscillator) circuits as well. Because of these variations in transistors and crystals, it is not unusual to find that a circuit we duplicate from an amateur journal does not perform as specified — or perhaps not at all! The author may have chosen the proper component values for his or her crystal and transistor, but they may be incorrect for your components.

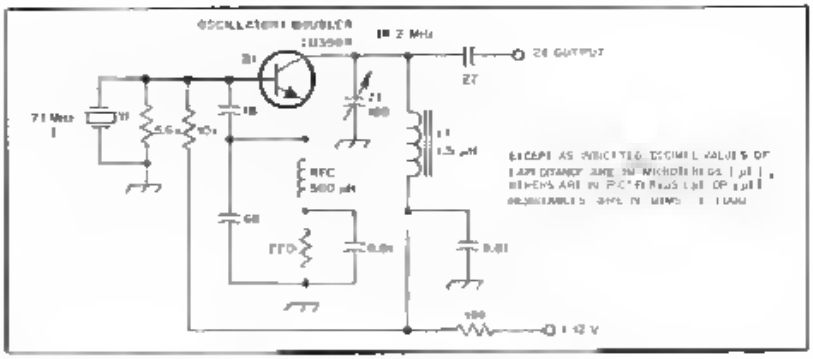


Fig. 2 — A crystal oscillator may be used as a frequency multiplier as shown here. The collector circuit is tuned to the desired harmonic of the crystal.

Oscillators That Change the Frequency

Earlier, we touched upon an oscillator that serves also as a frequency multiplier. We would not want to use such a circuit in a one-transistor QRP transmitter, but we could use it to drive a straight-through amplifier in a low-power transmitter. The reason we should avoid oscillator/multipliers directly into an antenna is because they are quite inefficient, and they would cause subharmonic energy to be radiated. The exception in the case of subharmonic radiation would be when we use well-designed filters in the transmitter output. The filters would have to reject the *oscillator* frequency as well as harmonics of the desired output frequency. This would call for a quality band-pass filter rather than the customary low-pass filter. In other words, we would need to reject frequencies above *and* below the desired output frequency.

An example of an oscillator/multiplier is shown in Fig. 2. The basic circuit is a Colpitts oscillator of the kind we saw in Fig. 1B. The difference is found in the collector circuit. C1 and L1 comprise a tuned collector tank that is adjusted for resonance at twice (2f) the crystal frequency. The collector of Q1 is no longer cold at RF, but has 14.2-MHz RF current present. Output from this circuit will be lower than that from the oscillator of Fig. 1B. This is because the efficiency of any multiplier is lower than that for a straight-through amplifier. Most oscillator/doublers exhibit an efficiency of approximately 33% after being optimized. Were we to triple or quadruple in the collector circuit (which is entirely acceptable), the efficiency would be correspondingly lower. The technique is useful when we are willing to amplify the oscillator/multiplier output by means of straight-through amplifiers. Generally, the CW note will be less prone to chirp if we multiply in the oscillator or in the stage immediately after the oscillator. For the most part, our cost will be minimal when we add an amplifier after a frequency-multiplier stage. Transistors and resistors are quite inexpensive these days!

Aids to Frequency Stability

Voltage regulation is important in an oscillator if the main power supply is not regulated. How can we achieve oscillator regulation simply and at low cost? A Zener diode regulator is the answer. The circuit of Fig. 3 illustrates the simplicity of Zener diode dc regulation. D1 is a 9.1-V, 400-mW regulator. It will hold the oscillator base and collector voltage constant during key-down conditions. R1 is the dropping resistor for the diode. Without this resistor, the diode would draw excessive current and burn up. If the resistor has too much resistance, the diode will not regulate at 9.1 V. Information on selecting the correct value of resistance is presented in the ARRL *Handbook*. We need to recognize that the Zener diode must draw a certain amount of current if it is to provide regulation. This current can range from 10 to

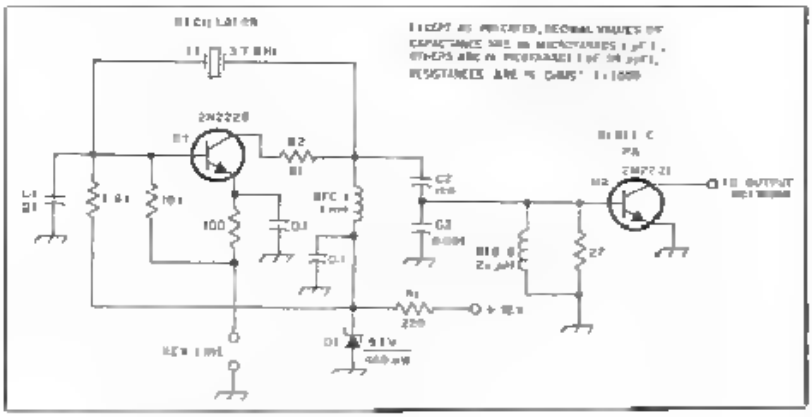


Fig 3 — An example of how a Zener diode (D1) can be used to ensure stable operating voltage (see text).

Table 1
Values of L and C for the One-Stage QRP Transmitter (Fig. 4)

Band	C1 (pF)	C_t (pF)	C2 (pF)	C3 (pF)	C4 (pF)	L1 (μH)
3.5 MHz	1000	330	100	1200	100	4.2, 29 turns no. 26 wire on T50-2 toroid core
7.0 MHz	560	180	100	600	47	2.12, 20 turns no. 26 wire on T50-2 toroid core
10.1 MHz	410	100	100	430	33	1.48, 19 turns no. 26 wire on T50-6 toroid core
14.0 MHz	412	68	100	300	22	1.0, 13 turns no. 26 wire on T50-6 toroid core

Note: C2 is a Mouser no. 24AA034 (page 78) 10-mm-diameter trimmer, 65-180 pF. C3 and C_t should be polystyrene or silver mica. All others are disc ceramic.

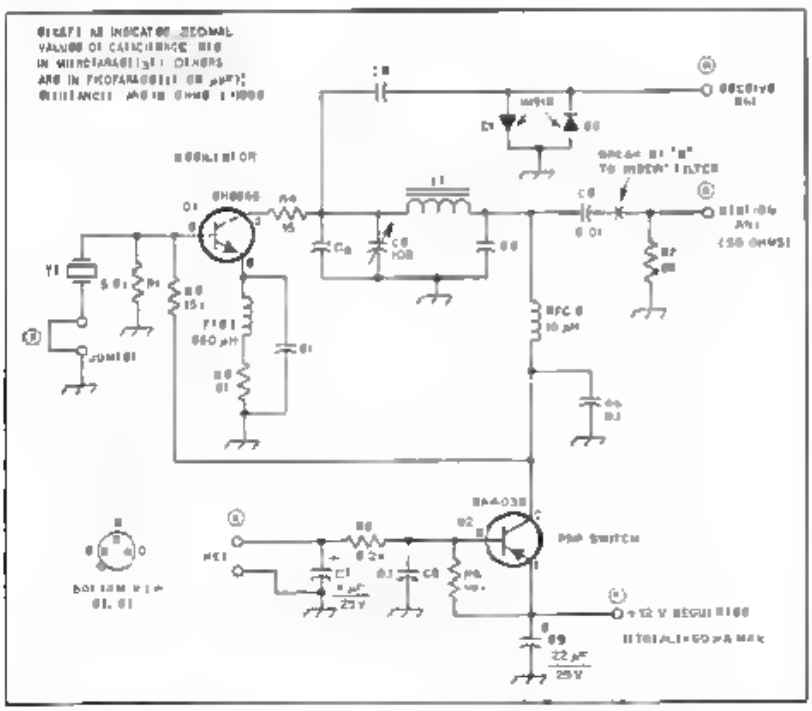


Fig 4 — Schematic diagram of a practical QRP transmitter. Resistors are ¼-W carbon composition. Fixed-value capacitors are disc ceramic. Capacitors with polarity marked are electrolytic or tantalum. C2 is a ceramic trimmer capacitor. See Table 1 for coil and capacitor data for the bands from 80 through 20 meters. Y1 is an International Crystal Mfg. Co. type GR, 20-pF load capacitance. Other fundamental crystals will be suitable (see text).

15 mA in this type of circuit, depending on the value of R1. For portable operation from a battery power supply, therefore, this added current must be taken into account when considering the discharge rate of the battery.

You will notice also that we have added R2 in the circuit of Fig. 3. This 22-ohm resistor is located close to the collector of Q1 to prevent unwanted parasitic oscillations that may occur at VHF. It is not unusual for a high-frequency transistor to self-oscillate in the VHF range when the circuit board is laid out with long copper elements. R2 serves as an inexpensive preventive device. Parasitic oscillation can impair the efficiency of the oscillator, encourage spurious radiation, and cause the CW note to be unclean. R2 may not be necessary in circuits that are laid out carefully.

Our circuit in Fig. 3 is a Pierce oscillator. C1 is a feedback capacitor, and C2-C3 in series provide the remaining part of the feedback divider. The effective value of C2 and C3 in series is 107 pF. These two capacitors function as an impedance-transformation circuit as well. The impedance at the top end of RFC1 is quite high compared to that of the base of Q2. The Q1 collector impedance is stepped down by virtue of the ratio of C2 and C3. The *ARRL Electronics Data Book* contains the equations and examples for using capacitive dividers to transform one impedance to another.

Simple Transmitter Performance

Some builders of homemade QRP transmitters are unmindful of the importance of proper impedance matching and output network design. It is not unusual to find some hams using one- or two-stage QRP rigs with output tank circuits that follow vacuum-tube concepts. That is, a single tuned output circuit is used, without regard for the collector and load impedance. Maximum power transfer can't be had without suitable matching of the impedances. Under some conditions of mismatch, the PA (power amplifier) stage may break into self-oscillation. This can cause spurious radiation, and it may even destroy the PA transmitter.

Take, for example, a tube QRP rig that has 150 plate volts and draws 10 mA when operating at 1.5-W dc input power. The plate impedance is 15,000 ohms. Conversely, a 1.5-W solid-state final amplifier that uses a 12-V collector supply will have a collector impedance on the order of 96 ohms. Attempts to use a tube type of output tuned circuit will be met with dismal results when dealing with a 96-ohm collector impedance! Furthermore, the transistor collector will load the high-impedance tank circuit and destroy the Q. This will permit harmonics to be radiated from the antenna, thereby causing TVI and interference to other services. The transmitting chapter of the ARRL *Handbook* contains detailed data

on impedance-matching networks for solid-state transmitters.

A Practical One-Stage QRP Transmitter

An uncomplicated circuit for QRP transmitting from 80 through 20 meters (see Table 1) is shown in schematic form in Fig. 4. DC power input is 250 mW (¼ W), which is ample for worldwide communications under good band conditions if an effective antenna is used (beam antenna, vertical radiator or dipole high above ground). This transmitter permits full break-in (QSK) without the use of antenna relays. The receiver antenna-input line is simply connected to point R of Fig. 4. When the key is up, the receiver is effectively attached to the station antenna. Upon closure of the key, the antenna line to the receiver is shorted to ground by means of D1 and D2. This transmit-receive (TR) circuit permits instant changeover from transmit to receive.

Q2 is a dc switch that serves as a keying transistor. When the terminals at K of Fig. 4 are shorted by the key or keyer, Q2 saturates and supplies +12 V to Q1, thereby turning on the oscillator. C2 is adjusted for maximum power output (100 mW), consistent with a chirp-free CW note. (The best note should occur at the dip in Q1 collector current — approximately 30 mA of total transmitter current.)

R5, C7 and C8 form a shaping network to soften the trailing edge of the CW waveform. The purpose of R7 is to provide a light load at the output of C5 under all conditions. I observed that the transmitter tended to be unstable before R7 was added, particularly when the SWR was higher than 2:1. The instability was observed while I was adjusting the Transmatch for an all-band antenna. Stability was fine when a low SWR was present while using a dipole fed with 52-ohm coaxial cable.

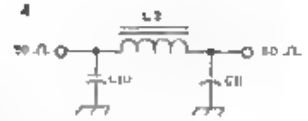
VXO Operation

The pair of terminals marked with an X in Fig. 4 identify the location of a jumper wire that can be removed to permit variable crystal oscillator (VXO) operation through insertion of a coil and variable capacitor. This modification is shown in Fig. 5. C_{VXO} can be a broadcast-band variable with half of the rotor plates removed; a 100- or 140-pF miniature variable capacitor may be used.

L_{VXO} is a small inductor wound on a toroid core. The greater the inductance, the more you will be able to shift the crystal frequency. Too large an inductance, however, will cause Q1 to function as a VFO rather than as a crystal oscillator. This will cause instability and chirp. Typical frequency swings are 3 kHz at 80 meters, 6 kHz at 40 meters, 9 kHz at 30 meters and 12 kHz at 20 meters. Experimental values for L_{VXO} are 12 μH (20 meters), 15 μH (30 meters), 20 μH (40 meters) and 30 μH (80 meters). Experiment with the number of coil turns until you obtain frequency shifts on the order of those listed.

Table 2
Low-Pass Filter Components

Band (Meters)	C10, C11 (pF)	L2 (μH)	Filter Cutoff (MHz)
20	220	0.5 13 turns no. 24 enam. on Amidon T37-6 toroid	15
30	270	0.72 18 turns no. 24 enam. on T37-6 toroid	13
40	390	1.0 15 turns no. 24 enam on Amidon T37-2 toroid	8
80	220	2.0 22 turns no. 26 enam. on T37-2 toroid	4



Values for use in building a single-section harmonic filter that can be added at the output of the QRP transmitter. C10 and C11 can be disc-ceramic, silver-mica or polystyrene capacitors.

The recommended crystals for VXO and standard use in this transmitter are plated AT-cut fundamental crystals in HC-6/U holders. A load capacitance of 20 pF is suitable for the crystals. Surplus crystals in FT-243 holders may not offer good activity, and they probably won't do too well for VXO operation.

Construction Notes

Keep the leads of all components as short as possible when soldering them to the PC board. Also, when winding L1, be sure to spread the coil turns around 2/3 of the toroid core. Bunching them too close together will increase the inductance, and spreading them over all of the core will decrease the inductance.

You may wish to add a single pi-section harmonic filter to the transmitter output in the interest of spectral purity. I have used this circuit with and without the filter, and have found the output (without the filter) to be clean enough to prevent TVI or harmonic radiation that could be detected on the air. Table 2 contains a circuit that can be added. The photograph of the transmitter shows the three added components for the filter. I made six holes with a no. 60 drill and soldered the filter in place. If you choose to follow this approach it will be necessary to sever the circuit-board foil between C5 and the antenna output terminal. A Mote Tool® or knife blade will be suitable for cutting the copper. Alternatively, you may mount the parts on a piece of perforated board and glue the subassembly to the main circuit board near the antenna output terminal. Spectral photos of the transmitter output before and after the

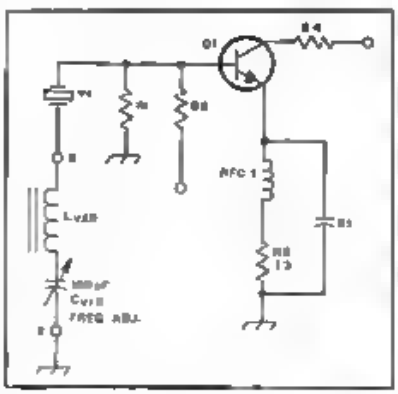
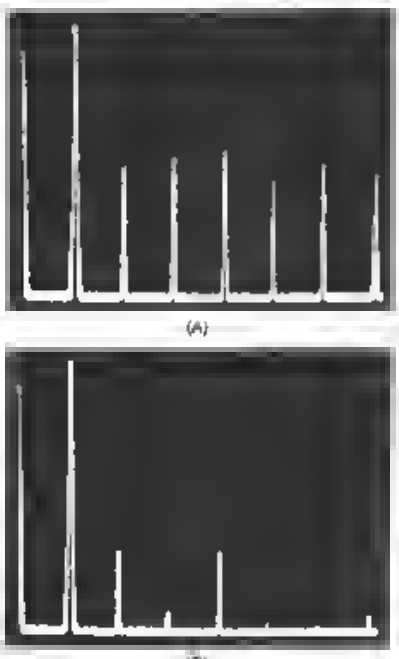


Fig. 5 — The jumper across terminals X of Fig. 4 may be removed to add these components. A coil (L_{VXO}) and a tunlue capacitor (C_{VXO}) are used to shift the crystal operating frequency (see text).



(A)

(B)

Fig. 6 — Spectral display of the QRP transmitter, without the filter (A) and with the filter (B). Horizontal divisions are each 10 MHz, vertical divisions are each 10 dB. Power output is approximately 0.25 W on 20 meters. At A, all spurious output is at least 34 dB down from peak fundamental output, at B it is least 52 dB down.

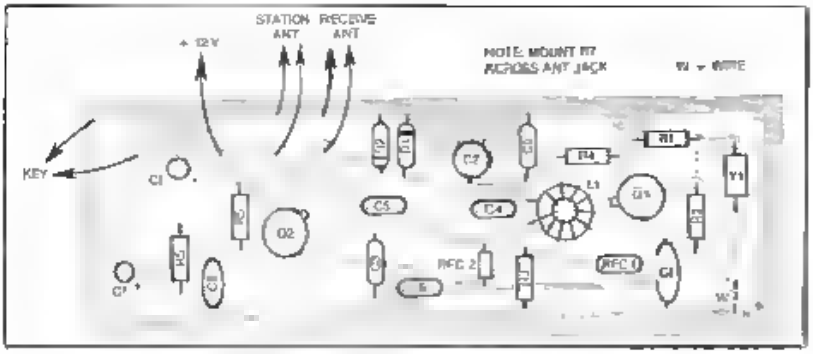


Fig 7 Component side view of the PC board showing placement of the components

addition of the simple filter are shown in Fig 6.

The completed assembly can be mounted in a cabinet or box of your choice. A nice homemade box can be fashioned from pieces of double-sided PC board that are soldered together where the sections join. The circuit need not be completely enclosed. Rather, you may prefer to mount the board on short standoff posts on an L-shaped piece of aluminum stock. The vertical part of the L can then serve as a panel for the jacks and VXO tuning capacitor. Adhesive-backed plastic feet can be affixed to the bottom of the L chassis. Your 12-V power supply can be packaged in the same box that contains the transmitter. A parts placement guide is provided in Fig. 7. A scale etching template is given in the Hints and Kinks column.

Adjustment and Use

Attach a 56-ohm resistor at the transmitter output to serve as a dummy load. Apply operating voltage and plug in your key. Hold the key down and tune your receiver to the transmitter frequency. Send some CW and monitor the note. If it is chirpy, adjust C2 until the CW note sounds proper. You can observe the S meter on your receiver while tuning C2 for maximum power output, consistent with a good-sounding CW note.

With an antenna connected (it should have an impedance of 50 ohms), look for a clear frequency and call CQ. It is wise to have two or three crystals available if you do not use VXO control. Don't despair if you don't receive an answer on the first few calls. That can happen even when running QRO (high power)! Eventually you will receive a response to your CQ, and the fun will commence. When answering someone on or near your crystal frequencies, try to respond to loud signals. This will mean that your signal will probably be fairly loud in the other station's receiver, assuming the operator is not running high power.

In Summary

If you haven't had the courage to work with transistors, this article may be the stimulant you've needed. On the other hand, if you've been building simple QRP rigs and have had poor results, the design tips we've discussed may get you headed down the right path.

There are many QRP operators in the world, so why not join them and face the exciting challenge of low-power operation. If you want to use this little transmitter in the field, take along a 12-V motorcycle battery, a lantern battery or 10 size-D cells connected in series. Of course, you will need a battery-operated receiver to use with your transmitter for field work. Numerous circuits for QRP receivers are described in the ARRL book, *Solid State Design for the Radio Amateur*.

Notes

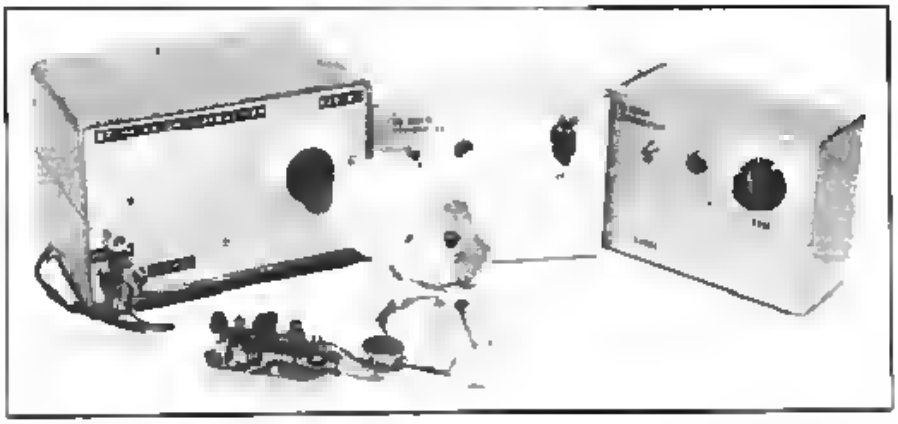
A "lid" is a poor operator, or one with a bad-sounding signal. The term comes from the early crystal Amateur Radio and is rumored to have been inspired by a bad CW signal that sounded like the lid on a kettle of boiling water.

Deleted.

Circuit board etching pattern for the QRP transmitter. The pattern is shown full size from the foil side of the board. Black areas represent unetched copper foil.

From December 1986 QST, p 19:

Three Fine Mice—MOuSeFET CW Transmitters



Got a hankerin' to build a simple CW transmitter that's a real performer? Take your pick—one or all—for 80, 40 and 30 meters. They're VFO controlled, too!

By Michael J. Masterson, WN2A
7 Hudson Rd
Budd Lake, NJ 07828

The availability of low-cost power MOSFETS (they're not really MOuSeFETs) creates new possibilities for "homebrew" transmitter design. For several years, I have used various RF bipolar transistors, all priced in the $12 to $16 range, in homebrewed CW transmitters. They performed well, but when I found inexpensive switching MOSFETS priced at $1 to $2, I thought, "Why not give them a try?" The low-cost devices seemed to be a painless (to the pocketbook) way to determine the MOSFET's potential and how to tame it for use at RF. Since I'd used a few RF bipolar transistors while learning how to use them, I knew certainly that a few MOSFETS would be "cooked" before I found the right circuit. I selected 30 meters as the test band, and eventually built transmitters for 30 and 40 meters.

Technical Approach

From articles I reviewed, it was apparent that the majority of MOSFET transmitter circuits use a 24-V, or greater, drain potential and most use RF-characterized device types. Because RF power MOSFETs are priced too high for this learning effort, my objective was to obtain at least 10 W output from a switching MOSFET operating from a 12-V supply.

Different circuits employing heavy gate swamping, RF feedback, drain loading and even the common-drain configuration were tried.

So, I did a lot of computer modeling, experimentation and article review. Different circuits employing heavy gate swamping, RF feedback, drain loading and even the common-drain configuration were tried. Some of these circuits looked promising—for a while. But, just as a circuit seemed to provide sufficient gain, the device would be destroyed by gate breakdown, not excessive dissipation or thermal runaway. I discovered that some form of gate protection is required, along with capacitive drain loading. Apart from the final amplifier, the remainder of the transmitter uses a proven transistor lineup from a previous design.

Circuit Description

Fig 1 is the schematic diagram of the transmitter. Frequency dependent parts information is given in Tables 1 and 2. The transmitter power chain is straightforward and is divided between two boards. (All transmitters use the same PC boards.) The VFO board contains Q1, an FET VFO, buffer Q2 and the balanced doubler composed of Q3 and Q4. D4 provides power-supply regulation for Q1. The 80-meter transmitter uses a Hartley VFO; it's a simple circuit and keeps the inductance of L1 at a reasonable value. A series-tuned Clapp oscillator with the inductor wound on an air-core ceramic form (for stability) is used in the 30- and 40-meter transmitters. Balanced doubler Q3 Q4 gets its drive from the bifilar winding on T1. The transistor collectors are tapped down on T2 for optimal output. T2 is tuned to the operating frequency, twice the VFO frequency.

The VFO runs continuously. When

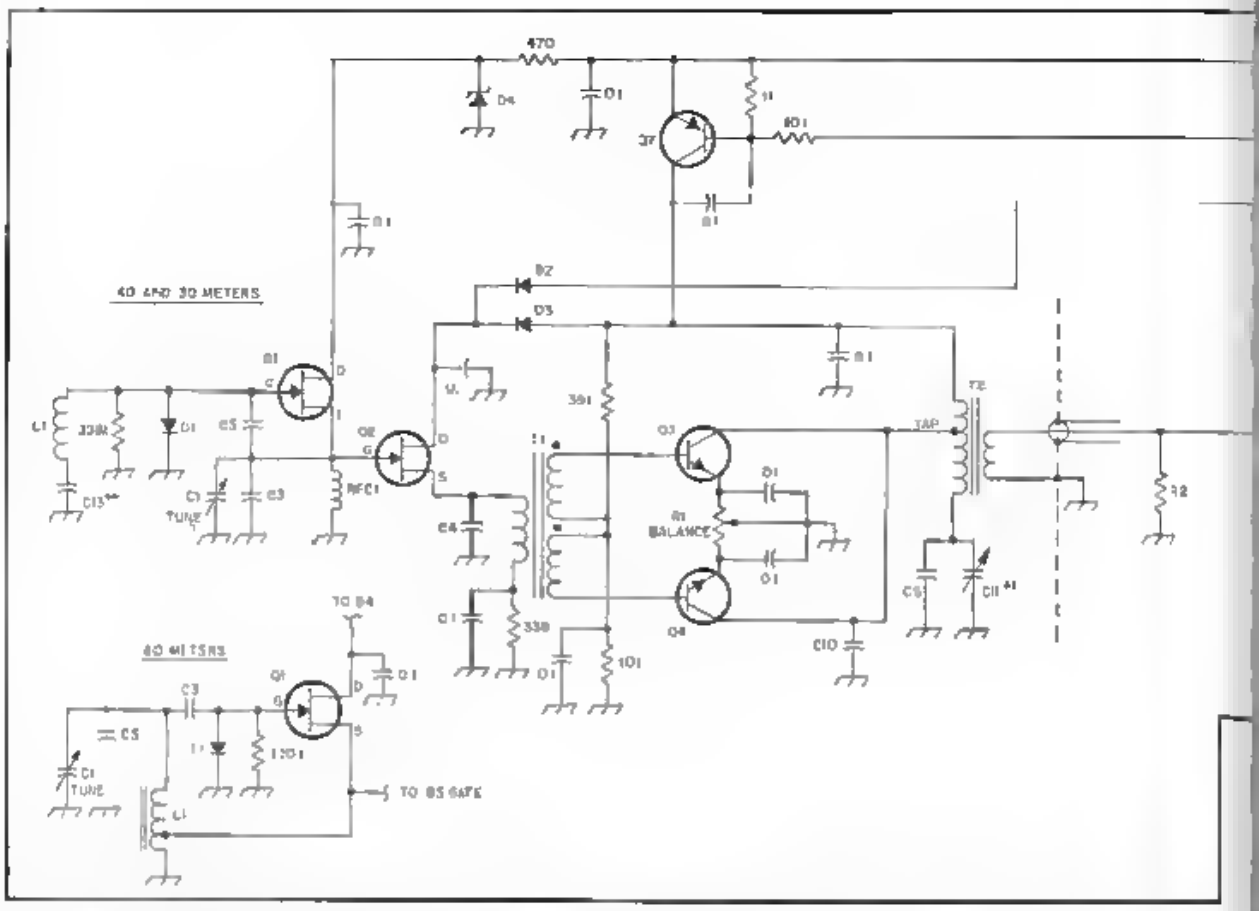


Fig 1—Schematic diagram of the MOSFET low-power transmitter. The 80-meter VFO is shown in the inset drawing. Refer to the parts list and Tables 1 and 2 for parts values not shown on the diagram. All 0.1-μF capacitors are 25-V X7R or Z5U ceramic types. All resistors are ¼- or ½-W, 10% tolerance acceptable.

C11—47-μF, 16-V electrolytic or tantalum.
C12—Deleted.
D1-D3 incl—1N4148.
D4—9.2-V, 400-mW Zener diode (1N959 or equiv).
D5—13-V, 400-mW Zener diode (1N964 or equiv).
FB—2t no. 28 enameled wire on FB-43-101 ferrite bead.
J1—Amphenol 126-011 jack (or equiv).
J2—Phono jack.
P1—Amphenol 126-010 plug (or equiv) for power cable.
Q1, Q2—MPF102 (see text).
Q3, Q4—2N3904 (see text).
Q5—2N3053 (see text).
Q6—80M, IRF 523; 40 m and 30 m, IRF510.
Q7—2N3906 or 2N2907A.
R1—500-Ω, 1-turn trimmer potentiometer.
RFC1—100-μH RF choke on phenolic form (Miller 4642 or equiv).

spotting, the buffer and doubler stages are keyed. The driver and final amplifier stages are on only during transmit when +12 V is applied at J1-E through an external TR switch. C12 provides some keyed wave-shaping, and R1 is an adjustment for doubler balance. C10 ensures stability at this stage, but it may not always be necessary. Measured VFO-board output is in the order of 60 mW. Improved keyed-wave shaping using a time-delay circuit was devised by Zachary Lau, KH6CP, of the ARRL Lab. That circuit employs a general-purpose PNP transistor, Q7, to moderate the otherwise fast rise time of the keyed wave. This addition also allows one side of the key to be grounded.

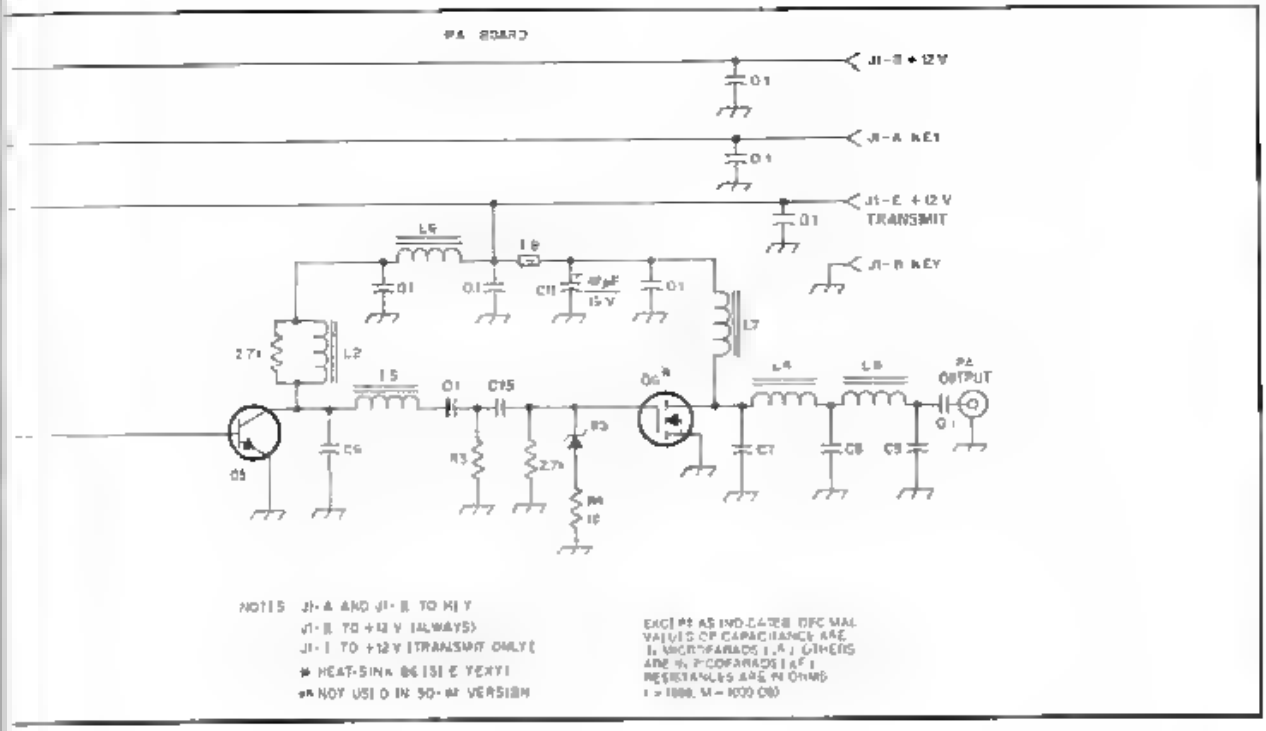
Q5, a 2N3053 operating class C, and Q6, an International Rectifier IRF type MOSFET also operating class C, comprise the power-amplifier chain. Q5 delivers about 1 W through an L network to Q6's gate circuit.

A power MOSFET's gate circuit is quite different from a bipolar power amplifier's base circuit. D5 provides two functions: it protects Q6 from excessive gate voltages and acts with C15 to provide a "grid-leak" action. During the negative half-cycle, D5 conducts and charges C15; during the positive half-cycle, C15's charge is added to the RF drive to supply a maximum of 15 V gate potential. Power MOSFETS have a high input resistance, but do require drive during switching. This is because of gate-source and gate-drain (Miller) capacitances. Gate-loading resistor R3, and drain-loading capacitor C7, augment stability as verified on a computer-aided design program. R4 limits the power dissipated in D5.

L4, C8, L5 and C9 form an L-pi output network (a pi-L in reverse). The L section matches the 5-ohm drain impedance of Q6 to a 100-ohm image impedance with a Q_L of 4.35. A pi network with a Q_L of approximately 2 then takes the 100-ohm image impedance down to 50 ohms. This type of network is less critical to tune than a T network for a given amount of harmonic suppression. At a nominal power output of 16 W, the second harmonic is 45 dB down; other harmonics are at least 60 dB down. (This performance was verified in the ARRL Lab.) The power level you obtain may be somewhat greater on 40 meters and less on 30 meters.

Component Notes

Most components used in this project are



Widely available from radio component stores, hamfests or mail-order distributors. A & A Engineering is a one-stop source for boards and parts for this project.[3] The devices recommended for use at Q1 and Q2 are not difficult to find, and a 2N4416 or 2N3823 may also be used. Q3 and Q4 should be matched for current gain, or at least originate from the same production lot. A match of 50% or better (at I_c = 10 mA) will suffice. A 2N2222A can be used at Q3 and Q4, and candidates for Q5 include the 2N2102 or D42C4; good results were obtained with the D42C4 at 80 meters. I purchased the MOSFET (Q6) from Frank, K2AW, at a hamfest table.[3] Motorola, RCA, GE, GI and other manufacturers make IR equivalents.

The toroidal inductors are available from Amidon or Radiokit.[3,4] All coils are wound with no. 28 enameled wire. After the VFO is built and tested (see tune-up and operation), hold the L1 windings in place with a thin layer of glue. L4 and L5 are wound with two paralleled wires to effectively increase the wire size and reduce copper losses. (These are *not* bifilar windings! You may optionally use a larger wire size instead.) Except where noted, capacitors are X7R or Z5U ceramic types. These are used for bypassing and decoupling functions, but not in tuned circuits. C2 and C3 are specified as NP0 ceramic units for excellent temperature stability. C4 through C10, inclusive, can be polystyrene, silver-mica or NP0 ceramics. Do not use X7R or Z5U ceramic capacitors here—degraded performance can result. Polystyrene capacitors work well in this circuit, and are compact compared to mica or NP0 ceramic types. Use caution when soldering polystyrene capacitors because excessive heat will melt the plastic.

Construction

The transmitter may be housed in any

[3]Notes appear at end of article

Table 1

Frequency Dependent Capacitor and Resistor Values

Capacitor Number	80 M	40 M	30 M
C1	25-pF air var	35-pF air var	35-pF air var
C2	450 pF (9 × 50 pF N)	1000 pF P	940 pF (2 × 470 pF P)
C3	50 pF N	470 pF P	600 pF (6 × 100 pF N/P)
C4	200 pF (2 × 100 pF N/P)	100 pF N/P	100 pF N/P
C5	200 pF (2 × 100 pF N/P)	100 pF N/P	50 pF N/P
C6	1000 pF P	470 pF P	330 pF P
C7	3300 pF P	1000 pF P	400 pF (4 × 100 pF N/P)
C8	2700 pF P	1410 pF (3 × 470 pF P)	1000 pF (10 × 100 pF N/P)
C9	1100 pF P	700 pF (7 × 100 pF N/P)	400 pF (4 × 100 pF N/P)
C10	50 pF N/P	50 pF N/P	
C13	Not used	250 pF (5 × 50 pF N)	250 pF (5 × 50 pF N)
C14	Not used	60-pF trimmer	60-pF trimmer

N = NP0 ceramic, P = polystyrene, N/P = NP0 ceramic or polystyrene. Silver-mica capacitors can be substituted for the polystyrene types.

Resistor Number	80 M	40 M	30 M
R2	47	68	68
R3	22	33	33
R4	10	10	10

sturdy, shielded enclosure. An 8 × 5 × 5-inch (LWD) cabinet provides more than enough room, even a 7 × 5 × 3-inch box, such as the LMB 782, is of ample size. The two circuit boards measure about 2 × 3 inches each, the VFO board being slightly larger than the PA board. A single-sided PC board (or perfboard and flea clips, if desired) is used for the VFO. This board can be installed in a box to shield it from the PA board, but no ill effects were found without the shield. If you elect to shield the VFO, C1 should be installed in the VFO box, and the larger transmitter cabinet used. Short, stiff leads from C1 to the VFO board are important for mechanical stability. Figs 2 through 5, inclusive, show the exterior and interior views of two of the three prototype transmitters. The 80-m transmitter (Figs 2 and 4) was the first unit built. Figs 3 and 5 are views of the 40-m unit.

The power amplifier PC board is double-sided, with one side left unetched. Copper foil or braid is used to provide low-inductance wraparounds to ground. Solder the foil to both sides of the board. Alternately, plated-through holes can be used in place of wraparounds. Solder all other components to this board prior to installing Q6. Cut the leads of Q6 to a length of 0.3 inch, and mount the transistor at the board edge. Insulating hardware for Q6 consists of a kapton or mica insulator, nylon washer and spaghetti sleeving for the metal bolt. Heat-sink grease is thinly applied to both sides of the insulator. (Too much grease impairs heat transfer.) The PA board and Q6 are bolted to the front panel, which acts as a heat sink. The leads of Q6 must not be stressed, so shim the PA board with metal washers if necessary. A small, clip-on heat sink will help Q5 dissipate heat.

Table 2
Frequency Dependent Inductance Values

Inductor Number	*80 M*	*40 M*	*30 M*
L1	14.5 μH 68t on T50-6, tap at 14t	14 t μH 40t no. 36 enameled wire on 3/8-in-dia ceramic form	7 μH 35t on 3/8-in-dia ceramic form
L2	13t on FT-37-61	9t on FT-37-61	9t on FT-37-61
L3	19t on T50-2 (1.8 μH)	12t on T50-6 (0.5 μH)	9t on T50-6 (0.33 μH)
L4[1]	15t on T50-6 (0.9 μH)	10t on T50-6 (0.43 μH)	8t on T50-6 (0.3 μH)
L5[1]	22t on T50-2 (2.6 μH)	15t on T50-2 (1.2 μH)	13t on T50-6 (0.9 μH)
L6	11t on FT-37-61	10t on FT-37-61	9t on FT-37-61
L7	11t on FT-37-61	9t on FT-37-61	6t on FT-37-61
T1	pri: 18t on FT-50-61 sec: 9 bifilar turns	pri: 16t on FT-50-61 sec: 8 bifilar turns	pri: 12t on FT-50-61 sec: 10 bifilar turns
T2	pri: 40t, tap at 20t[2] sec: 7t on T50-2	pri: 26t tap at 11t[2] sec: 5t on T50-2	pri: 22t, tap at 10t[2] sec: 4t on T50-2

All inductors wound with no. 28 enameled wire unless otherwise noted.
[1]L4 and L5 are wound with two parallel lengths of no. 28 enameled wire (this is done to increase the effective wire size. These are *not* bifilar windings.
[2]Tap measured from Q7 side of primary.
Powdered iron toroids (T50-6 and T50-2) and ferrite toroids (FT-37-61 and FT-50-61) are available from A & A Engineering, Amidon Associates or Radiokit (see notes 1, 3, 4). Ferroxcube 3/8-in 4C4 and 1/2-in 4C4 ferrite toroids may be substituted.

Tune-Up and Check-Out

With the exceptions of C1 and C14 (if used), there are no other variable capacitors in the transmitters. All tune-up is done by adding or removing turns on the toroidal inductors, and by compressing or expanding the windings. This may take some time and patience, but it results in compact construction without the need for large trimmer capacitors. Start with one or two extra turns on L1, the primary of T2, L3, L4 and L5, and remove turns as required during tune-up.

First, adjust the VFO tuning range by listening to its output with a calibrated receiver or coupling a frequency counter to the VFO output. (If your frequency counter is not sensitive enough, you'll have to use a receiver initially.) Set R1 at mid-range and apply +12 V to J1-D, with a key across

Fig 2—An 80-m transmitter was the first unit built. The enclosure was salvaged from a piece of defunct test equipment.

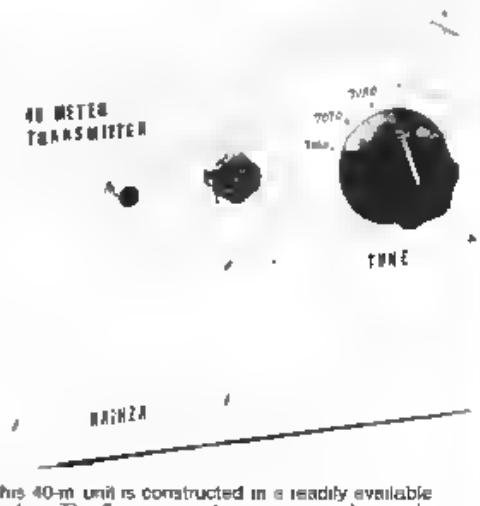


Fig 3—This 40-m unit is constructed in a readily available aluminum box. The 5-pin connector serves as a key and power-input jack.

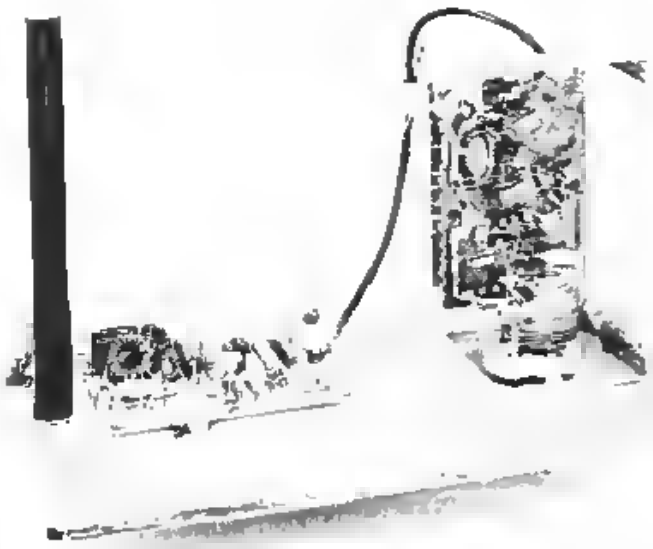


Fig 4—A close look at this inside view of the 80-m transmitter reveals the VFO is built on perf board. Note the shielded VFO enclosure (cover removed). To the left is the driver/output board. Q6 may be seen in the foreground attached to the heavy front panel.

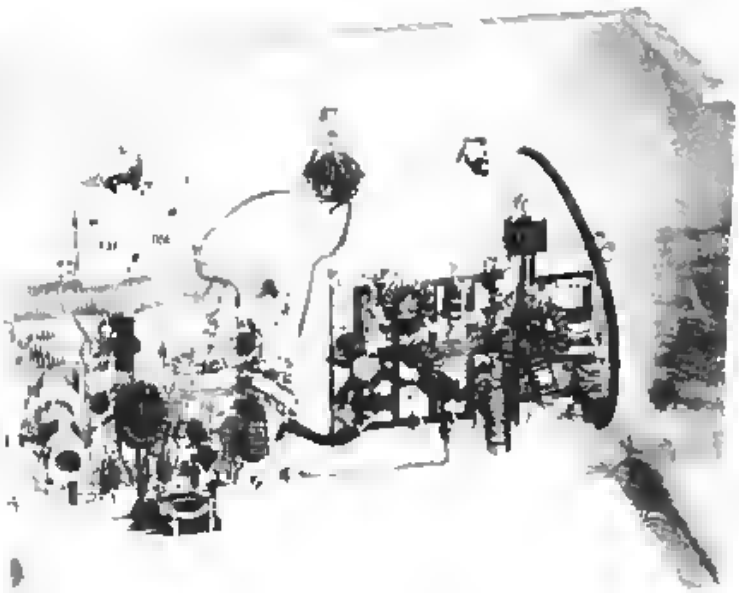


Fig 5—An inside view of the 40-m unit. PC boards from A & A Engineering were used in this model.

J1-A and J1-B. Adjust L1 for the desired band coverage, as you vary C1. Next, set C1 to mid-band and adjust the primary of T2 (80 meters) or C14 (30 and 40 meters) for maximum output as indicated on a sensitive power meter connected to VFO output. If C14 does not allow you to tune through a maximum output point within its range, adjust the primary of T2 until it does. Using a wave or dip meter, sense around T2, and set R1 for minimum fundamental feedthrough (that is, at 3.8 MHz on 80 meters, 3.5 MHz on 40 meters and 5 MHz at 30 meters). You should notice a setting of R1 where the fundamental nulls out. Then, disconnect the power meter and connect the VFO output to the PA input using a short length of coaxial cable.

Next, connect an RF power meter to J2 and apply +12 V to J1-D and J1-E. Keying the transmitter briefly, adjust L3, L4 and L5 for maximum output. Go back and

. . . you should have 12 to 20 W of RF output

adjust the primary of T2 (80 m) or C14 (30 and 40 meters) to peak the output reading. Again, adjust L3, L4 and L5 if necessary to maximize output power. At this point, you should have 12 to 20 W of RF output. (With or without parts substitutions, your results may vary from mine because of construction or other differences. Prior RF-circuit building experience should help you correct any problems.) Finally, adjust R1 for minimum fundamental feedthrough as heard on a receiver. Again a null should be found. Set the receiver to the transmitter's output frequency and reduce the receiver RF gain. Key the transmitter. The transmitted note should sound clean, with no chirp or clicks. Check the heat dissipation of Q6; if it is too hot to touch, it may not be heat-sinked properly. No stability problems were noted in my units; however, a check-up on a spectrum analyzer would help determine if any excessive spurious signals exist. I performed the tune-up as described here without the use of a spectrum analyzer, but if you have access to one, use it!

On The Air

Operation is simple. Use a Transmatch and a resonant antenna. I recommend that you use a 12-V regulated power supply capable of delivering 2.5 A. In my receiver, provision is made for off-the-air monitoring. During receive, key the VFO to spot your operating frequency. External TR switching should remove the voltage from J1-E during receive. While transmitting, monitor your off-the-air signal instead of using a sidetone. No hint of thermal runaway has been noted, and the transmitter sustained no damage with high SWR loads. With the values of C1 given,

No hint of thermal runaway has been noted, and the transmitter sustained no damage with high SWR loads.

frequency coverage is about 100 kHz on 80 m, 60 kHz on 40 m, and all of 30 m.

Summary

The on-the-air performance of these little rigs is quite satisfactory. Using a folded dipole on 30 meters, TK5, IV3, G, F, FG and North America have been worked. Results on 80 m (using a random-length wire antenna) are good from Southeast to Midwest states and Canada. I have been too busy (and having fun) building these rigs to get on 40 m, so it is up to you to find out how one of these MOuSeFET transmitters will perform on that band! Though they're small, they pack quite a bite!

Acknowledgments

I offer my sincere thanks to Herb Englemann, W2VIE, and Mike Kucks, KA2ZAM, of KDI Electronics for use of lab facilities; the use of the KDI Electronics facilities was invaluable to the design effort. My thanks also to my wife, Dawn, for her encouragement during this project.

Notes

[1]A & A Engineering, 2521 W LaPalma Ave, Unit K, Anaheim, CA 92801, tel 714-952-2114

[2]K2AW's "Silicon Alley" 175 Friends Ln, Westbury, NY 11590

[3]Amidon Associates, 12033 Otsego St, North Hollywood, CA 91607, at 213-780-4429

[4]Radiokit, PO Box 411, Greenville, NH 03048, tel 603-878-1033

For updated supplier addresses, see ARRL Parts Supplier List in Chapter 2.

From May 1979 QST, p 16:

Transmitter Design — Emphasis on Anatomy

Part 1: Which is best — duplication of a published circuit or an understanding of how the circuit works? This builders course provides some "hows" and "whys" for a 10- to 15-watt, 40- and 20-meter cw transmitter.

By Doug DeMaw, W1FB
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A heap of burned-out transistors, some unsavory language and a hastily scrawled sign which read, "Help Stamp Out Transistors," greeted me as I walked into a friend's workshop recently. Fred stood there with a deeply furrowed brow and pointed to a wretched-looking, pc-board assembly which had been worked and reworked until it looked like no hope remained for it. Fred is one of those fellows who loves to build amateur gear, but never took the time to change his thinking from vacuum tubes to semiconductors. He could duplicate the circuits in amateur magazines, but couldn't make them "play" when something went amiss. After some casual conversation and a hot cup of coffee, Fred calmed down and we began troubleshooting his problem child. The major faults were instability in the PA stage and low output from the driver. An hour later we had his transmitter percolating nicely, and Fred poked his thumbs proudly into his chest and proclaimed, "Ain't it a beaut?"

It occurred to me as I sensed my friend's anguish that a better understanding of how a solid-state circuit functions would have saved him countless hours and a considerable amount of grace in the eyes of The Almighty. The foul language and extra money spent for transistor replacements could easily have been avoided. My adventures with Fred helped to inspire this course in transmitter anatomy. Knowing why a particular circuit was chosen by the designer, and how it is supposed to function in the composite assembly, should help you avoid the "Freddie syndrome."

Understanding Our Circuit

The circuit for our workshop project was based on numerous requests for a transmitter that would serve as a mate for "The Mini-Miser's Dream Receiver" which appeared in *QST* for September, 1976. A power output in the 10- to 15-watt bracket seemed suitable for most of the QRP applications one might encounter, and ample power would be available for driving an amplifier later on should the builder be motivated toward QRO.

Fig. 1 shows the block diagram of the transmitter. Let's run through it and see what each section does. Starting at the left we find a 7-MHz VFO. It operates straight through on 40 meters. The arrows show that S1A/S1B route the rf energy directly to the broadband amplifier module during 7-MHz operation. For use on 20 meters, the VFO output is switched to a push-push doubler by means of S1. Output at 14 MHz is applied to the broadband amplifier when the switch is set for

Two versions of the 7- and 14-MHz cw transmitter are shown here. At the left is the W1FB prototype. On the right is a model built by WA8UZO. Both units are small and lightweight.

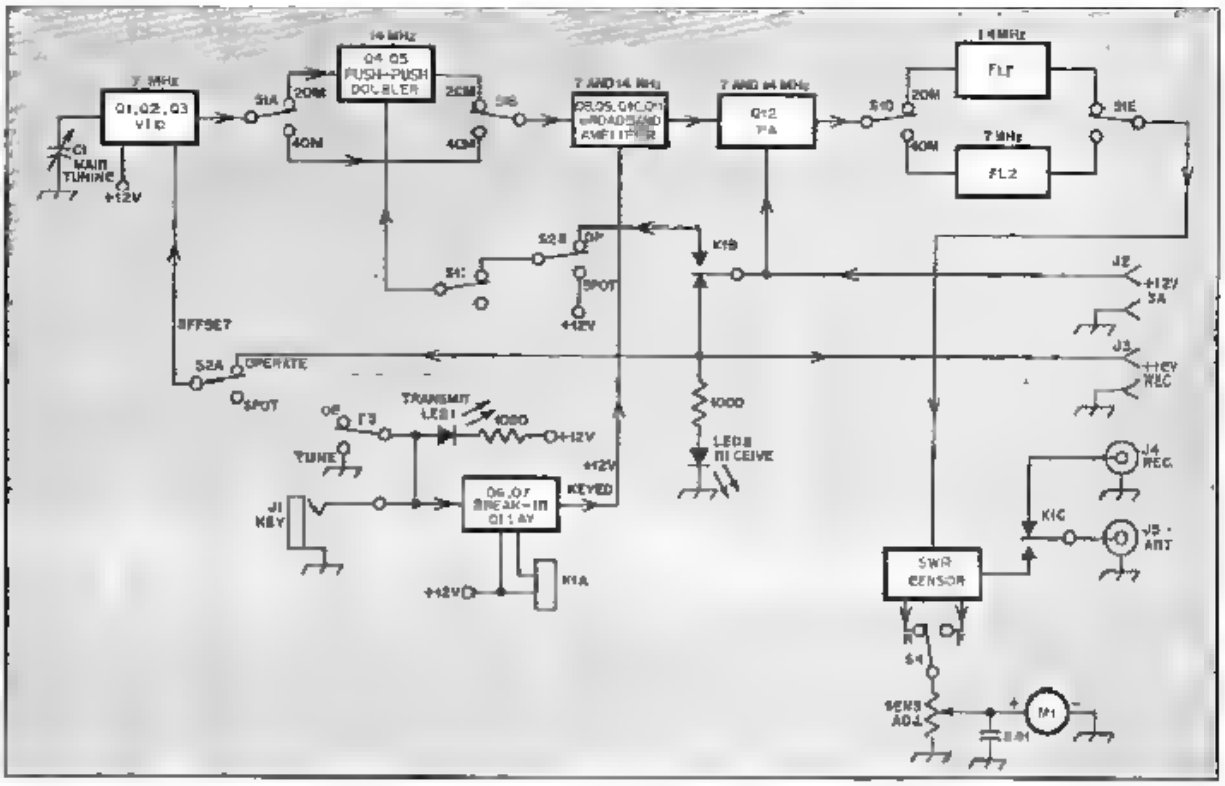


Fig 1 — Block diagram of the two-band, solid-state transmitter. The arrow heads indicate the direction of signal or dc voltage flow. Coaxial cable (miniature) should be used between the various modules when completing signal paths. S1 is a five pole, two-position, wafer switch with phenolic insulation. S2 is a dpdt miniature toggle switch. S3 and S4 are spst miniature toggle switches.

20-meter operation. You will notice that an offset line goes to the VFO. When S2A is in the OPERATE position and the key (J1) is open, relay contacts at K1B place +12 V on the VFO-offset line. This voltage turns on a switching diode in the VFO. The diode switches some additional capacitance into the VFO tuned circuit and moves the operating frequency outside the amateur band. This prevents an unwanted beat note in the receiver tuning range during the receive period. When the transmitter is keyed the offset voltage is disconnected by means of K1B, and the VFO provides output on the desired operating frequency. It is necessary to disable the offset circuit for spotting (zero beating), so S2A is placed in the SPOT position for that function. Operating voltage must be applied to the push-push doubler during 20-meter spotting, and S2B is used for that purpose. Activating the doubler assures a loud beat note when zero beating another 20-meter signal.

As the signal moves to the right in Fig 1 it reaches the broadband amplifier. This circuit was chosen because it requires no tuned circuits. Elimination of tuned, narrow-band circuits at the output of each of the three amplifiers in the module makes it possible to avoid complicated band-switching circuits. The broadband amplifier delivers approximately 1 watt of output and requires only 10 mW of rf energy from the VFO or doubler to develop its rated output power. Actually, the broadband amplifier is useful from 1.8 to 30 MHz, even though this transmitter covers only two bands. The amplifier is biased for Class A (linear) operation so that it can be driven easily by the VFO. The linearity is not a necessary feature for cw use, however, but would be ideal if this were an ssb exciter.

To the right of the broadband amplifier is a PA stage. It is driven to a power output of 10 to 15 watts by the 1-watt signal from the previous module. A Motorola MRF449A transistor is used in the PA. It is capable of 30 watts of output, and has a rated gain (typical) of 13 dB at 30 MHz. Our purpose in restricting the output to 15 watts is to minimize the overall current drain of the transmitter to 3 amperes or less. This will assure longer battery life during portable operation, and will simplify the requirements of an ac-operated dc supply (regulated). The actual amount of rf output power will depend upon the characteristics of the last stage in the broadband amplifier and the PA transistor. This results from the slight nonuniformity in transistor manufacture: Some have more gain than others. It is for this reason that an output figure of 10 to 15 watts is given.

At the far upper right of the block diagram are two filters — one for each band. They are selected by means of S1D/S1E. Since the PA is also a broadband amplifier there will be a substantial amount of harmonic current in the output. To keep the unwanted energy suppressed by 40 dB or greater it is necessary to use FL1 and FL2. The filters are low-pass type (T networks). They are pretuned, so no external peaking controls are needed.

Output from the filters is routed through an SWR-sensor circuit (lower right of drawing). A panel meter, M1, serves as a visual indicator for trimming an antenna or adjusting a Transmatch for a low SWR. The latter is essential if proper operation of the PA stage is to be realized. Relay contacts at K1C transfer the antenna from the transmitter to the receiver during standby periods.

At the lower left of Fig 1 we have a break-in delay module. It has a variable time constant which controls the drop out time of the changeover relay, K1A. The amount of delay time can be determined by adjustment of a potentiometer on the circuit board. Closure of the key charges the timing capacitor, which in turn actuates a bipolar-transistor dc switch. The switch closes K1A and applies operating voltage to the broadband amplifier. S3

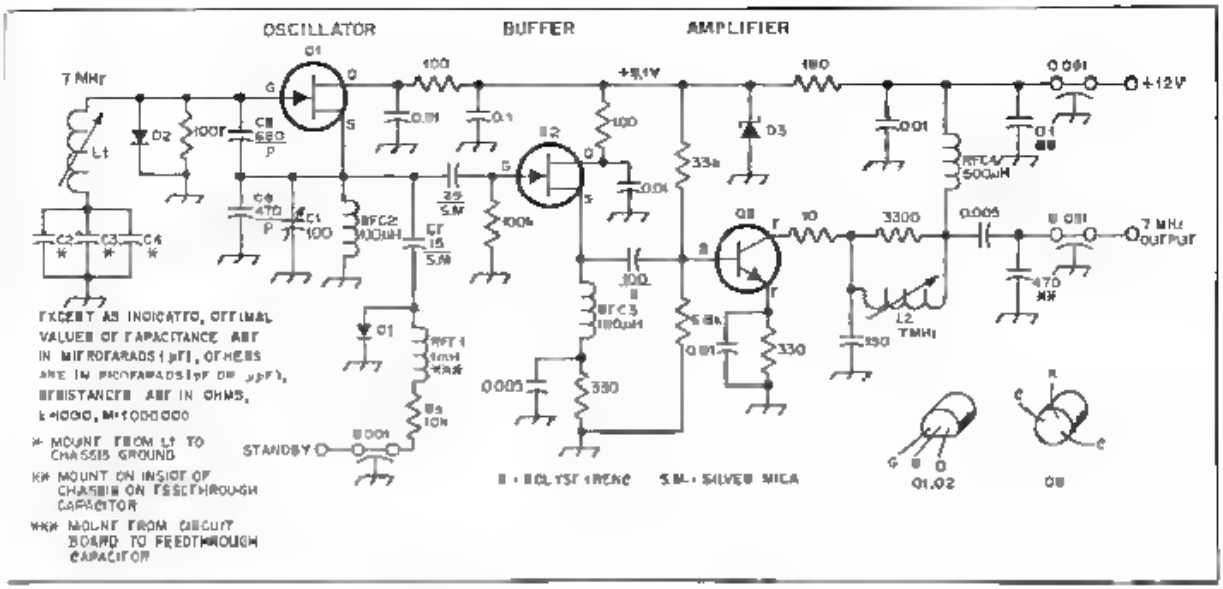


Fig. 2 — Schematic diagram of the VFO. Fixed-value capacitors are disk or chip ceramic unless otherwise indicated. Resistors can be 1/4- or 1/2-W composition. (See June, 1976 *ham radio* for the author's detailed explanation of this VFO circuit.)

C1 — 100 pF, miniature air variable.
C2 — 50 pF, polystyrene.
C3 — 10 pF, silver mica.
C4 — 27 pF, polystyrene.
D1, D2 — High-speed silicon diode, 1N914 or equiv.
D3 — 8.1-V, 400-mW or greater Zener diode; 1N5239 or equiv.
L1 — Slug-tuned inductor with 6-µH nominal inductance (Miller 42A686CBI in VA1F8 unit, Miller 23A475RPC in WA8UZO model).
L2 — Slug-tuned, pc-board mount inductor, 3.2-µH nominal inductance (Miller 23A476RPC or 25 turns no. 32 enam. wire close-wound on Miller 27A014-6 form).
Q1, Q2 — Vhf JFET, MPF102, 2N5486, 2N4416 or HEP802.
Q3 — 2N2222A or HEP-53001.
RFC1-4 — Miniature rf choke (Millen J30t or equiv.).

locks the break-in delay circuit into the key-down mode for tune-up purposes. An LED indicator illuminates during transmit periods, and a second LED indicates when the circuit is in the standby (receive) mode. At that time the transfer relay routes 12 volts to the receiver via J3. This control voltage can be used for muting and unmuting the receiver.

Understanding the VFO

The VFO of Fig. 2 has a familiar face, as it has been used in a number of my circuits.[1] It has been such a faithful and predictable performer that it was chosen again. The circuit at Q1 is a Colpitts oscillator, but some of you may prefer to call it a series-tuned Clapp if you date back to the tube era when that type of circuit emerged as one of the more stable varieties of VFO.

Three capacitors (C2, C3 and C4) are used in series with L1 to ground. This method permits a larger amount of inductance to be used at L1 than would be possible in a more common, parallel-tuned, VFO tank. The higher inductance is less subject to changes in value from heating than would be the case if high C and low L were used. Three capacitors are used below the coil rather than one so that the circulating rf current will be divided among them. This lowers the heating in any one capacitor and improves stability.

C5 and C6 are feedback capacitors that take part of the oscillator output (source terminal) and route it back to the input (gate). This feedback is what causes the FET to oscillate. RFC2 is used to keep the feedback energy at the source of Q1 while providing a dc return to ground for the FET. Stated simply, it's an isolating choke for the rf.

Another purpose is served by C5 and C6: They add a considerable amount of shunt capacitance from the FET base to ground. This helps to disguise the small changes in FET junction capacitance during operation — a significant contribution to oscillator stability. D2 gets into this act, also. It conducts on the positive swing of the oscillator rf voltage, and that limits the change in FET junction capacitance. (Maximum capacitance change occurs near the peak of the positive half of the sine wave.) In addition to helping stabilize the oscillator, D2 reduces the harmonic output of Q1. This is because nonlinear changes in junction capacitance encourage the generation of harmonic currents. It is necessary to use a high-speed, rf type of diode for this purpose, such as a 1N914 switching kind.

C7, D1 and RFC1 are used in the VFO-offset circuit. When the +12 volts are applied to D1 as discussed earlier, C7 is placed in parallel with the main tuning capacitor, C1. This moves the VFO operating frequency lower so that the signal won't be heard in the receiver during standby. R_s is used to prevent damage to the diode; it limits the current through the diode junction when the offset voltage is applied through it and RFC1.

The 0.01-µF capacitor and 100-ohm resistor at the drain of Q1 are used to place the drain at ac ground (bypass) and to isolate Q1 from the other transistors in the VFO module. This is called a decoupling network, and it helps prevent unwanted self-oscillation in the remaining VFO-drain stages. Q2 has a similar decoupling network in the drain circuit.

A buffer stage (Q2) is shown in Fig. 2. It functions as an isolation circuit between the oscillator and Q3. It is used as a source follower — the output being taken from the source element of the FET. Because the gate of an FET has a very high impedance (megohms), the transistor does not load the output of Q1. The gate coupling capacitor is small in value (39 pF), and that also reduces the loading effects on Q1. The lighter the loading, the less chance there will be for oscillator "pulling" (chirps) when the transmitter is keyed. Because Q2 is a source follower it will not provide a voltage gain. Actually, a slight loss will occur at Q2. Typically, a voltage gain of 0.9 will be realized when using this type of buffer stage. This means that we lose 10 percent of the rf voltage that is applied to the gate of Q2.

RFC3 is used as a broadly resonant (low-Q) tuned circuit that peaks at 7 MHz with the approximate 5 pF of stray circuit

[1]Notes appear at end of article.

capacitance. Zener diode D3 is used to obtain a 9.1-volt regulated supply for Q1 and Q2. This prevents changes in oscillator frequency when the 12-volt power supply output changes. Regulated voltage is supplied to Q2 so that it maintains relatively constant operating characteristics. Voltage shifts at Q2 could cause slight changes in internal capacitance and resistance, and those variations could cause some pulling of the oscillator.

VFO Output Stage

It will be necessary to have ample drive to the broadband amplifier stage of Fig. 1. VFO buffer Q2 could not provide sufficient excitation to operate the remainder of the transmitter. Therefore, we have added Q3 to build up the VFO output power. This amplifier stage operates in Class A and uses a high-frequency, bipolar transistor — a 2N2222A. A 10-ohm resistor is placed near the collector terminal to discourage vhf parasitic oscillations. At 7 MHz the resistor offers minor resistance to the signal, but at vhf it looks like a high impedance, this prevents parasitics.

A pi network is used as the output tank for Q3. It is a low-pass type of network, which means it will attenuate harmonic energy. A 3300 ohm resistor is used in parallel with L2 to broaden the response. This will assure relatively constant VFO output to provide an even drive across all of the 40- and 20-meter cw bands.

The output capacitance for the pi network is obtained by utilizing the capacitance of the feedthrough terminal (C3) and the 470-pF shunt capacitor. The collector tank is designed to transform the 500-ohm output impedance at Q3 to 50 ohms at the pi-network output. Even though the input impedance of the first stage of the broadband amplifier is on the order of 500 ohms, this mismatch is desirable. The lower the VFO output impedance, the less chance there will be for pulling effects caused by the later stages in a transmitter. The base-bias voltage for Q3 is taken from the 9.1-volt regulated line to further reduce the chance for pulling at Q1.

Assembling the VFO

Double-sided pc board material is used as a shield box for the VFO. Fig. 3 shows the pc-board pattern and includes a parts-placement guide. Ready-made pc boards or parts kits for the entire transmitter are available from a supplier.[1]

The components should be assembled on the etched circuit board before the side walls are soldered together around the VFO board. A pencil type of soldering iron with a fine tip is recommended for this and all other modules of the transmitter. Excessive heat will damage some of the components, and can cause the pc-board pads to come loose from the base material. Therefore, a 25- or 30-watt iron is the largest size that should be employed.

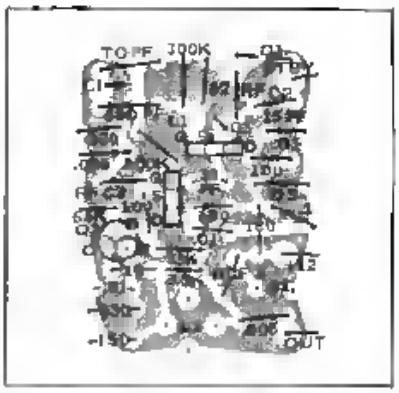
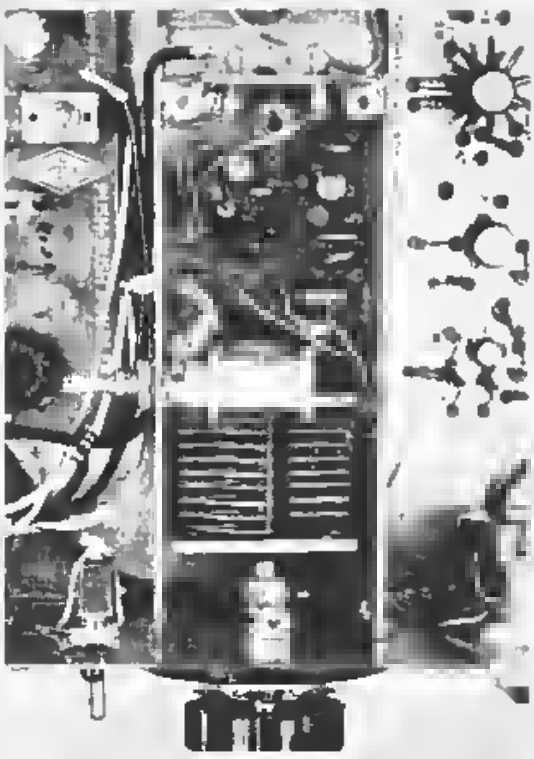


Fig. 3 — Scale layout of the VFO circuit board showing parts placement from the component side of the board.

Alignment

VFO testing can be accomplished by shunting the output to ground with a 560-ohm, 1/2-watt resistor and applying +12 volts where indicated on Fig. 2. Attach a two-foot piece of hookup wire to the output and place the loose end near the antenna terminal of a receiver. Next, set C1 so that the plates are fully meshed. With the receiver adjusted to receive 7.0 MHz, move the slug in L1 until the VFO signal is heard. At this point you can adjust L2 for maximum output at 7.1 MHz. The S meter on the receiver will be helpful when tweaking L2.

The offset circuit can be tested by connecting +12 volts to the offset line. The VFO signal can be expected to shift lower in frequency, as stated earlier. There should be no evidence of chirp when keying the 12-volt supply to the VFO.



A look into the VFO compartment of the WA@UZO model. Coil L1 is mounted on a side wall of the VFO box. Feedthrough capacitors are used as terminals for leads entering and leaving the VFO compartment.

Footnotes

DeMaw, "QRP Shakedown, Caymanian Style!" *QST*, March, 1975.

[1]Negatives, pc boards or complete parts kits for this project can be obtained from Bob Shriner, WA@UZO, Box 969, Pueblo CO 81002.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

From June 1978 QST, p 25:

Transmitter Design—Emphasis on Anatomy

Part 2: A VFO by itself doesn't offer much when it comes to transmitting, so let's proceed with the physical structure of our two-band transmitter. Here is some useful information on the frequency doubler and cw break-in delay circuits.

Perchance you're wondering why our VFO described earlier couldn't be made to operate on 14 MHz as well as on 7 MHz. Well, there's no reason why the L and C components couldn't be modified to provide two-band coverage. In such an example a band switch would be included in the VFO module for the purpose of selecting the 7- or 14-MHz coils and capacitors. The disadvantages of this scheme are at least twofold. Mechanical instability is likely to result from the switch contacts and related leads. Also, the effects of oscillator pulling are more pronounced as the operating frequency is increased. Concerning the latter, it would be a difficult task to prevent chirp during 20-meter cw work if the VFO were operated at 14 MHz.

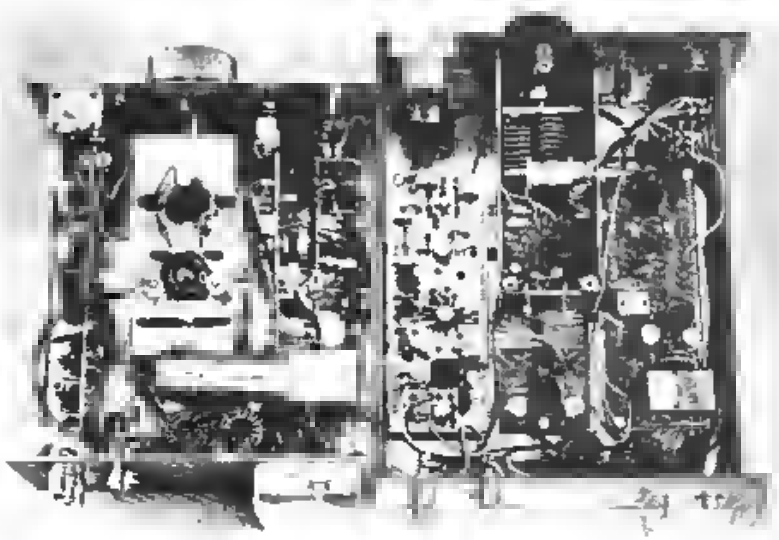
A more suitable technique at the higher operating frequencies is to employ the oscillator at one or more octaves below the desired excitation frequency, and utilize multiplication to obtain the required output frequency of the VFO chain. Through this process the mechanical instability is diminished greatly, and the frequency-multiplier stage or stages tend to isolate the oscillator from the load more effectively than would be the case with a straight-through buffer or amplifier.

Fig. 4 contains the circuit we will use for multiplication. Rather than follow the VFO chain with a single-ended frequency doubler (one transistor), we have elected to use what has long been known as a *push-push doubler*. Although the bases of the transistors are connected in push-pull by means of broadband transformer T1, the collectors are tied in parallel. In this manner the stage differs from a *push-pull* amplifier, as the latter would have the collectors as well as the bases in push-pull. A push-push amplifier favors *even* harmonics, whereas a push-pull amplifier does its best job with *odd* harmonics. Furthermore, a push-push doubler is practically as efficient as a straight-through amplifier. A single-ended doubler would exhibit a typical maximum efficiency of only 50 percent as opposed to a push-push doubler with a ball-park efficiency of 70 percent. There is no reason why a pair of JFETs couldn't be used at Q4 and Q5 of Fig. 4. If they were, however, the doubler output for this transmitter would be somewhat lower (inadequate) than with the 2N2222As we have employed.

Circuit Description

The VFO in Fig. 2 (Part 1, May QST) has a single-ended output terminal, so if we are to supply drive to the doubler of Fig. 4 it will be necessary to use a balun-

Interior views of the W1FB (left) and WA8UZO (right) versions of the transmitter. The push-push doubler/break-in delay module is at the far right in this photograph. The VFO and SWR-sensor modules are at the center of the WA8UZO unit, and the broadband amplifier may be seen at the left of the VFO. The PA stage is mounted on the rear wall (lower right) of each rig. The homemade heat sinks are visible on the back aprons of the enclosures.



type transformer (T1). The energy reaching the bases of Q4 and Q5 must be of opposite phase to assure push-pull drive to the doubler. To accomplish this we have included T1, a trifilar-wound broadband transformer (three wires wound on the core at the same time). The black dots on the schematic diagram, at the top of T1, identify the phase relationship of the windings. It can be seen that one transistor base is fed 180 degrees out of phase with the other, thereby satisfying our need for push-pull drive. Forward bias is supplied to the doubler stage through the junction (C and E) of the two right-hand windings. A 0.01-µF bypass capacitor brings that point in the circuit to rf ground.

For proper operation of a frequency multiplier it is necessary to establish Class C operating conditions. The forward bias on Q4 and Q5 implies Class AB operation, but the output from the main VFO chain overrides the forward bias and drives the doubler into the Class C mode. Bias is applied only to make the doubler easier to drive.

In the interest of optimum doubler performance it is necessary to establish dynamic balance. Most discrete transistors of a given type number exhibit different electrical characteristics. In our application we are concerned mainly with any difference in transistor gain which might exist. Ideally, Q4 and Q5 should perform in an identical manner. A balancing control, R1 in Fig. 4, has been included to enable us to match the operating traits of the two devices. A 47-ohm resistor is used on each side of the control to prevent the emitters from going directly to ground if the control arm is set at either end of its range. R1 is adjusted so that the output waveform (14 MHz) is as pure as possible. If R1 is set incorrectly there will be a substantial amount of the 7-MHz driving energy present at the collectors of Q4 and Q5. The worse the imbalance, the greater the level of the 7-MHz energy.

A tuned circuit (C8 and L3) is used at the doubler output to increase the available rf output voltage. A pure waveform would be attainable if only the 1000-ohm shunting resistor was used, but the doubler output would be quite low because of the dc voltage drop across the resistor. L3 permits the full supply voltage (less the drop across the 33-ohm dc coupling resistor) to reach the collectors of Q4 and Q5. Also, the 1000-ohm resistor broadens the tuned-circuit response to provide a nearly constant output level across the VFO tuning range. Fig. 5 provides the pc-board pattern and parts placement guide for the doubler and break-in delay circuits.

Fig. 4 — Schematic diagram of the push-push doubler. Fixed-value capacitors are disk ceramic. Resistors are 1/2 W composition, except for R1 (see below).
C8 — 510-pF mica compression trimmer (E-msnco 506 suitable).
L3 — Toroidal inductor. 17 trifilar turns of no. 26 enam. wire on a T50-2 powdered-iron toroid core.
R1 — Pc-board mount carbon control.
T1 — 17 bifilar turns of no. 28 enam. wire on an FT-50-61 ferrite toroid core. Twist wires approximately eight twists per inch before winding on core.

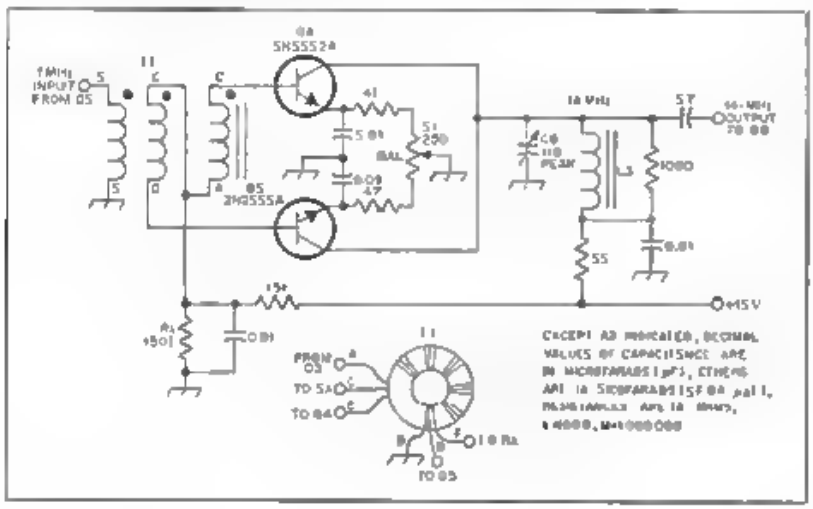
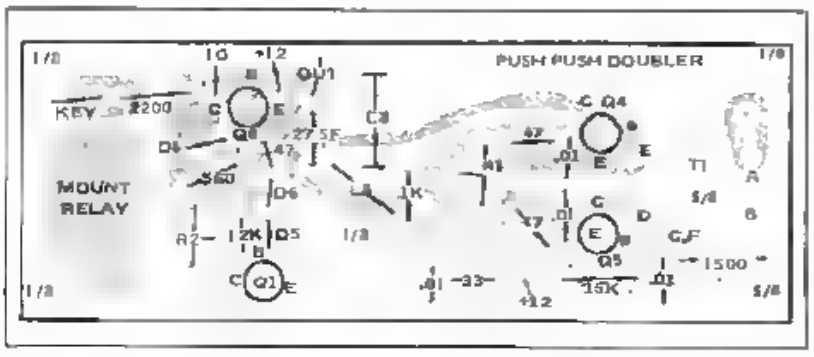


Fig. 5 — Parts layout and parts placement for the doubler/break-in delay board. View is from the component side of the board. 1/8 indicates a 1/8-inch hole.



The Final Touches

Checkout of the doubler is an easy assignment. The VFO module is connected to points A and B of T1. A 56-ohm resistor is attached temporarily between the doubler output (to the right of the 27-pF output coupling capacitor) and ground. The 56-ohm resistor simulates the load presented by the broadband amplifier (to be described later).

A short length of hookup wire is attached to the junction of the 56-ohm resistor and the 27-pF capacitor. The opposite end of the wire is placed near the antenna terminal of a receiver which is tuned to 7 MHz. Next, operating voltage is applied to the VFO chain and doubler. R1 and C8 can now be adjusted by setting them for *minimum* signal response at 7 MHz, as noted on the receiver S meter. If an oscilloscope is available, connect the scope probe to the top of the 56-ohm load resistor and adjust R1 and C8 for the purest waveform obtainable at 14 MHz. There may be some interaction between the adjustments of R1 and C8, so the foregoing steps should be repeated two or three times to ensure optimum doubler operation.

A low value of coupling capacitor (27 pF) is used to prevent the approximate 50-ohm input impedance of the broadband amplifier strip from loading C8 and L3 excessively. During 40-meter operation the push-push doubler is bypassed so that the VFO output goes directly to the broadband-amplifier module.

Break-In Delay Circuit

A cw break-in delay circuit is not an essential part of a transmitter, but it does provide an operating convenience which

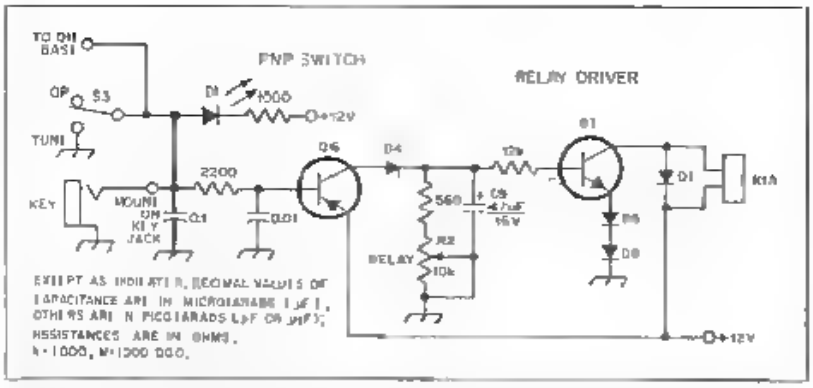


Fig 6 — Schematic diagram of the break-in delay circuit. Disk-ceramic capacitors are used except for C9 if it is electrolytic. Fixed-value resistors are 1/2-W composition. D4-D7 inclusive are 1N914 silicon diodes. K1 is a dpdt 12-volt dc relay. A Potter & Brumfield 24-V dc relay (unpins) was used in the W1FB version of the transmitter. The spring was stretched to lessen the tension, enabling the relay to close satisfactorily at voltages as low as 11. The relay number is KHP17D12. A 12-volt version is available. Q5 is a 2N3906 or HEP715. Q7 is a 2N1711 or HEP736. R2 is a pc board-mount carbon control.

makes it worth including. Manual switching could be used in place of the delay circuit to control the changeover relay, K1. Similarly, keying could be done by breaking the 12-volt supply to the keyed stages of the transmitter. The main advantage in utilizing a break-in delay system is that the operator has one less switch to manipulate between the transmit and receive modes. In some instances this permits faster mode changing than would be possible if mechanical switching was used.

Fig. 6 shows the break-in delay circuit. A straight key, bug or electronic keyer is connected to the input of Q6. When the circuit is completed, Q6 is effectively biased into conduction. At that time 12 volts appear at the Q6 collector to place a charge in C9. R2 is set for the desired discharge time of C9. The greater the resistance, the longer the period before C9 bleeds to ground. D4 is used in the collector lead of Q6 to serve as a one-way path for the dc voltage — a gate of sorts. This component was added after two transistors were destroyed at Q6 by a voltage peak which originated (after Q6) as the key was closed. D4 permits the +12 volts to flow into the charging network, but prevents positive-voltage transients from flowing back toward Q6.

When C9 is charged sufficiently to provide the forward bias necessary to turn on Q7, current flows through the field winding of K1, causing the relay contacts to close. As the voltage across C9 drops (key open), a point will be reached at which Q7 has insufficient forward bias to provide the collector current needed to keep K1 energized. At that time the relay will open. D7 is placed across the relay field coil to clip voltage spikes caused by the inductive "kick" when the relay field collapses. The spike, if great enough in amplitude, can travel along the 12-volt bus and damage transistors elsewhere in the transmitter. Damage could occur to Q7 as well.

The diodes in the emitter return of Q7 are used to establish approximately 1.4 volts of fixed-value bias for Q7. Depending on the transistor used as the relay driver, the resting current of Q7 may be high enough to keep K1 closed even though C9 has been nearly discharged. D5 and D6 prevent such an event from happening. The LED indicator used in parallel with the key was added by WA0UZO in his model of the transmitter, but it is not essential to the operation of the circuit. It illuminates when the key is closed, thus functioning as a *transmit* indicator (a full).

The break-in delay module can be tested by merely applying operating voltage and shorting from the key terminal to ground. If all is as it should be, K1 will close. R2 can be set for the delay time desired. If wiring errors have been avoided, and if no defective components were used, the "Freddie syndrome" should have remained dormant so far.

From July 1978 QST, p 23:

Transmitter Design — Emphasis on Anatomy

Part 3: Broadband power amplifiers eliminate the need for complicated band-switching circuits. Some amateurs believe that they are mysterious and hard to build. 'Tain't so!

It's unlikely that Freddie would have been able to design the broadband amplifier we are describing here, but he certainly should have enjoyed success in duplicating it and making it perform correctly. However, find something malfunctioned in his assembled module his chances of ocating the anomaly would have been enhanced greatly by an understanding of how a broadband amplifier functions. Let's consider the subject of how one of these critters does its particular "thing."

A broadband amplifier is intended to do precisely the job its name implies — amplify signal energy over a broad slice of the frequency spectrum. In meeting this requirement the amplifier should provide reasonably uniform output power across the band of frequencies it is designed to accommodate. Thus, if the circuit was designed to cover from, say, 3.5 to 14 MHz, and deliver 5 watts of output, there should be 5 watts of output available (no more and no less) at any discrete frequency within that range. In practice it is difficult to obtain that kind of precision, but a variation in power no greater than ±10 percent can be realized in a carefully designed amateur circuit.

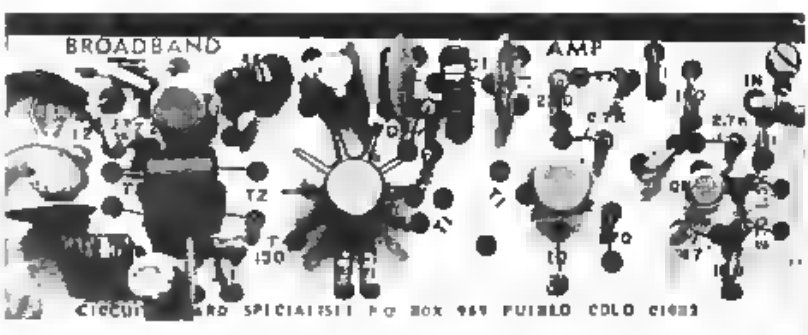
Solid-state amplifiers tend to supply increasing amounts of output power as the operating frequency is decreased. That is, a given transistor will exhibit more gain at 3.8 MHz than it will at 7 or 14 MHz. Therefore, in order to obtain a relatively flat frequency response from a solid-state, broadband amplifier it is necessary to use certain compensating elements to "taper" the overall gain downward toward the lower end of the amplifier operating range. The inclusion of feedback networks is the most common approach to this design criterion. The mathematical solutions to feedback design problems are beyond the scope of this article, but in-depth data on the subject are given in the ARRL book, *Solid State Design for the Radio Amateur*.

The required feedback for a broadband amplifier is usually introduced by means of R and C components between the collector and base of the transistor (negative feedback), and through the inclusion of degenerative feedback in the emitter circuit. Concerning the latter, the emitter bias resistor is bypassed for rf at the higher end of the amplifier frequency range (low-value capacitor), but is bypassed less effectively as the operating frequency is lowered. At the lowest end of the amplifier range the emitter may function as if no bypass capacitor was there at all. In ordinary language we are saying that the less effective the bypassing the lower will be the stage gain. This kind of frequency-response shaping can be further enhanced by selecting specific values of coupling capacitance between amplifier stages. That is, a low value of capacitance will be less effective as a coupling device at the low-frequency end of the range than it will at the high-frequency end of the range.

The feedback resistors and capacitors used between the collector and base of a broadband amplifier are chosen with the same design philosophy in mind. In this case the lower the operating frequency the greater the feedback voltage through a given value of base-to-collector resistor. The greater the feedback, the lower the stage gain. In cases where the feedback resistor is so low in value that excessive forward bias would reach the transistor base, a blocking capacitor is added in series with the resistor and forward bias is obtained by means of a separate resistive divider.

Broadband transformers are also used

Closeup view of the broadband linear amplifier



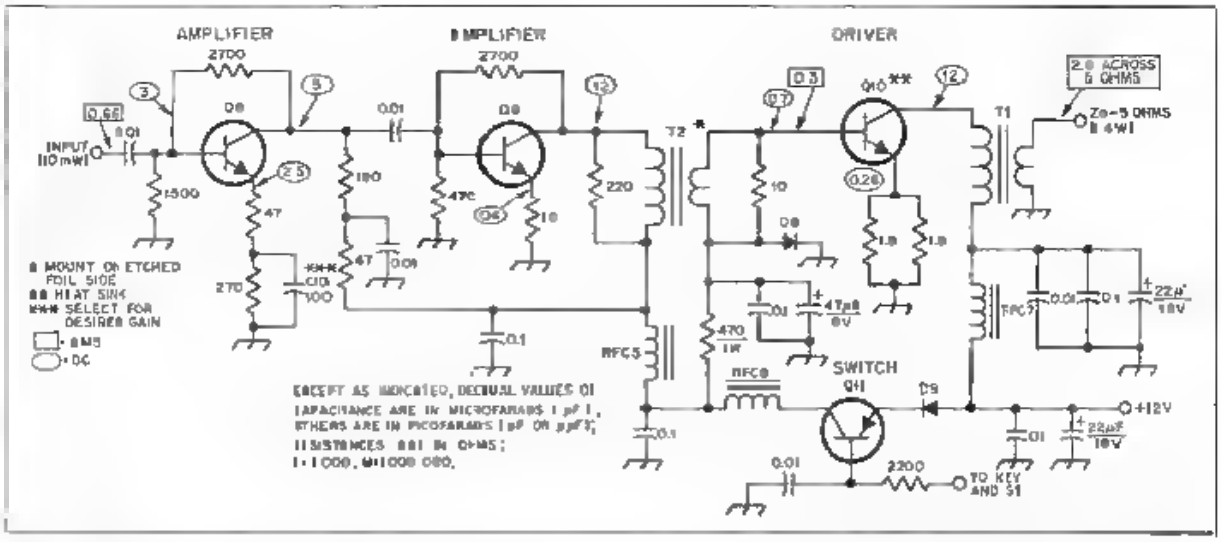


Fig 7 — Schematic diagram of the broadband linear amplifier. Capacitors are disk ceramic except those with polarity marked which are electrolytic or tantalum. Resistors are 1/2-W composition unless otherwise noted.

C10 — See text.
D8, D9 — 1-A 50-PRV silicon (1N4003 suitable).
Q8 — 2N2222A or equivalent.
Q9 — 2N3866 or HEP S3006.
Q10 — 2N2270 or HEP S3001.
Q11 — 2N4037 or HEP S3012.
RFC5-RFC7, incl. — 19 turns of no. 28 enam. wire on FT-37-43 ferrite toroid core.
T2 — Primary has 30 turns of no. 28 enam. wire on a FT-50-43 ferrite toroid core. Secondary has 4 turns of no. 28 enam. wire wound over cold end of primary winding.
T3 — Primary has 19 turns of no. 28 enamel wire looped through a BLN-43-302 ferrite core. Secondary has four turns of no. 28 enam. wire looped through the same core. Primary leads come out of end of core opposite the secondary leads.

in the type of amplifier under discussion. They are designed to operate as untuned rf transformers with a turns ratio chosen to match the output of the amplifier stage to its load (collector of one stage to the base of a succeeding stage, for example). A deliberate mismatch is sometimes introduced by the designer to achieve amplifier stability. Another approach is to shunt one or both of the transformer windings with a resistor. This tends to lower the transformer Q, which in turn discourages self-oscillation. The trade-off is in reduced stage gain.

Examination of Our Circuit

The broadband amplifier used in our transmitter is shown in Fig. 7. It was inspired by a similar circuit in the Atlas 210X transceiver. With approximately 10 mW of driving power at the input to Q8, the amplifier output at Q10 will be roughly 1.4 watts at 7 and 14 MHz. The input impedance of the composite amplifier is close to 50 ohms.

Feedback is provided at Q8 and Q9 by means of the 2700-ohm resistors connected between the collector and base of each stage. Degenerative feedback for Q8 is obtained by leaving part of the emitter-bias resistance unbypassed (47-ohm resistor). No bypassing is used across the 10-ohm emitter resistor of Q9. The parallel 1.8-ohm resistors in the emitter return of Q10 serve two purposes: They are unbypassed to provide degenerative feedback, and they help to protect the transistor from drawing excessive current (thermal runaway).

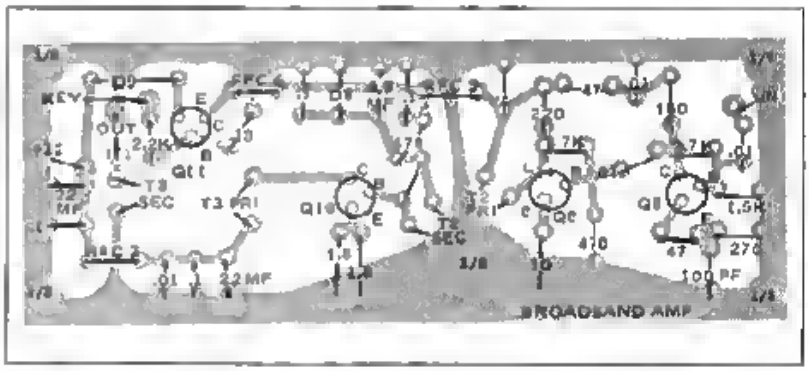


Fig 8 — Parts placement guide for the amplifier pc board. The shaded area represents an X-ray view of the etched foil pattern, this view is from the component side of the board. Same is used for this module. The foil on the component side is used as a solid groundplane, having only clearance holes for the mounting of the components. The commercially made board shown in the photo also has component identification information etched on this side of the board. All mounting holes on pc board marked with a "1/8" should be drilled through with a 1/8-inch drill bit. The board can then be mounted to the chassis with 6-32 machine screws, spacers and nuts.

T2 is a broadband toroidal-wound transformer. It is loaded on the primary by a 220-ohm resistor. A 10-ohm resistor is in parallel with the secondary winding. These resistors were added to reduce the drive to Q10, and to cure a low-level oscillation which occurred during the checkout period. T3 is also a broadband coupling transformer. It is wound on a ferrite core of the balun type. In the breadboard model of this amplifier an RCA 40082 transistor was used at Q10. Owing to its gain and f_T characteristics, it was somewhat more "lively" than the 2N2270 of Fig. 7. To obtain equal performance it was necessary to bridge the primary of T3 with a 150-ohm resistor. This ensured stability.

All three amplifiers are biased for linear operation (Class AB). This has no special value in a cw or fm transmitter, as Class C amplifiers are adequate for those modes. The primary advantage in using a linear amplifier in our transmitter is to lower the driving-power requirements (the transistors require less excitation voltage) and to lessen the occasion for harmonic generation in the stages (Class C amplifiers are richer in harmonic currents). The forward bias applied to Q10 is

developed across D8, which regulates the bias by virtue of its barrier voltage (0.7 volt for a silicon diode). A 470-ohm dropping resistor is used between D8 and the 12-volt supply bus to prevent the diode from consuming excessive current.

Decoupling networks are used in the 12-volt line between stages. This aids in preventing feedback (positive) from one stage to another. An excessive amount of feedback will cause self-oscillation of one or more of the stages. At Q8 a 47-ohm resistor and 0.01-μF capacitor comprise the decoupling circuit. RFC5, RFC6 and the two 0.1-μF bypass capacitors are used for this purpose at Q9. RFC7 and the related bypass capacitors are employed at Q10 to decouple the stage from the 12-volt line. High, medium and low values of capacitance are used at Q9 and Q10 to assure adequate decoupling at lf, hf and vhf. (The stages could self-oscillate at any of those frequencies.) Who needs or wants to be haunted by the "Freddie syndrome?"

A pnp bipolar switch (Q11) is shown in Fig. 7. It operates in the same manner as Q6 of Fig. 6. When the key is closed, Q11 conducts and permits +12 volts to reach Q9 and the bias network for Q10. A one-second oscillation occurred in the breadboard version of the transmitter, caused by the decoupling capacitors at Q9 and Q10. This formed a timing circuit which was triggered by a self-oscillation at Q11. The decoupling capacitors at Q9 and Q10 acted as a tuned-collector/tuned-emitter circuit for Q11. The oscillation caused the break-in-delay circuit to cycle at a one-second rate. This resulted in a repetitive cycling of the relay, K1. Insertion of D9 at Q11 cured the problem by providing a one-way gate in the feedback path. A crown type of heat sink is needed at Q10 to prevent damage to the transistor.

Amplifier Testing

Following completion of the assembly procedures given in Fig. 8, amplifier testing can be done. Tests can be performed first by connecting the VFO directly to the input of Q8 of Fig. 7 (40 meters). A 5-ohm, 2-watt load resistor should be attached across the secondary of T3. Apply operating voltage and short the keying line to ground. A VTVM and an rf probe can be used to compare the circuit voltages with those of Fig. 7. Approximately 2.6 volts rms will appear across the 5-ohm load resistor if the circuit is working correctly. If the overall amplifier gain is too low, increase the value of C10 experimentally. Although 100 pF was right for the circuits built by W1FB and WA6UZO, variations in transistor gain may require that less feedback be used at Q8. These tests can now be repeated at 20 meters, using the push-push doubler between the VFO and broadband amplifier.

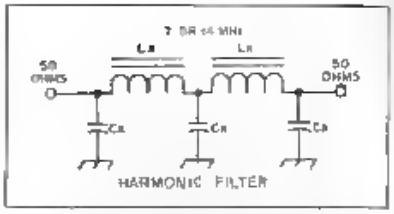


Fig. 8 — Diagram of the half-wave harmonic filter discussed in the text. For 7-MHz use L_a is 1.1 μH (15 turns no. 25 enam. wire on a T50-2 toroid core). C_a is 470 pF and C_b is 910 pF. For 28-meter operation L_a is 0.55 μH (10 turns no. 26 enam. wire on a T50-2 toroid core). C_a is 240 pF and C_b is 470 pF.

This much of the transmitter can be put on the air if the builder likes true QRP work, but it should *not* be connected to an antenna unless a harmonic filter is placed in the output line from T3. Furthermore, the turns ratio for T3 will need to be changed to provide a match to a 50-ohm filter and antenna. The secondary winding of T3 will require 15 turns rather than four turns if this is done. Fig. 9 gives the details for half-wave filters which can be used at 7 and 14 MHz, respectively.

From August 1978 QST, p 16:

Transmitter Design — Emphasis on Anatomy

Part 4: The final touches are applied to our transmitter by adding a 15-watt amplifier and an SWR indicator. If all goes well, we will become immune to the "Freddie syndrome"!

It is unlikely that the 1.5 watts of output from our broadband amplifier (Fig. 7) would lead to the acquisition of five-band DXCC. But a few more decibels might make such an endeavor a reasonable assignment. The amplifier described in this section will help, as the ew signal should be increased some 10 dB in strength!

The final-amplifier stage is shown in Fig. 10. A 220-ohm feedback resistor is used between the base and collector of Q12. An 1800-pF blocking capacitor has been included to prevent the collector dc voltage from being shorted to ground via T3 of Fig. 7.

This amplifier has an input impedance of approximately 5 ohms at 7 and 14 MHz. The 10-ohm base resistor is used as a preventive measure against instability, but only if needed. To remove some vhf harmonics which appeared at the collector of Q12, it was necessary to include the 330-pF bypass capacitor. At 7 and 14 MHz the capacitor has negligible effect on circuit performance.

As was the case with the stages in our broadband amplifier, decoupling of the 12-volt bus is necessary at Q12. This is accomplished by means of RFC9 and the related bypass capacitors. Once again, bypassing is done for lf, hf and vhf.

Since the amplifier is to operate in the Class C mode, no forward bias is used at the base of Q12. For all practical purposes, Q12 draws no current during key-up conditions. When drive is applied (key closed) the transistor is driven into the cutoff region to establish Class C operation.

The collector load impedance of Q12 is determined in the usual manner, where Zo = $Vcc^2/2Po$. Thus, for a 12-volt collector supply and a power output of 15 watts, we obtain a collector load of 4.8 ohms. T4 is a broadband transformer which is made from six toroid cores (see inset drawing of Fig. 9). It must transform the collector impedance to 50 ohms so that a suitable match and power transfer to the T-network filters can be obtained. A 3:1 turns ratio will suffice despite the slight mismatch (9:1 impedance ratio).

In order to prevent excessive harmonic energy from reaching the antenna it is necessary to include a filter at the output of Q12. FL1 and FL2 are used for this purpose. Each is a T type of low-pass network. Energy above the operating frequency is attenuated by the filters, but energy below the filter cutoff frequency passes without impairment. A spectral analysis of this transmitter indicated that all spurious output energy was at least 40 dB below peak power at the fundamental frequency. Additional attenuation could be realized by cascading two such filters at the PA output. The characteristic impedance of the filters in Fig. 10 is 50 ohms.

Fig. 10 — Circuit for the 10- to 15-watt Class C power amplifier. Capacitors are disk or chip ceramic unless otherwise noted. Capacitors with polarity marked are electrolytic or tantalum.

L6 — 3 turns no. 18 enam. wire on a T68-6 toroid core.

L5 — 10 turns no. 18 enam. wire on a T68-6 toroid core.

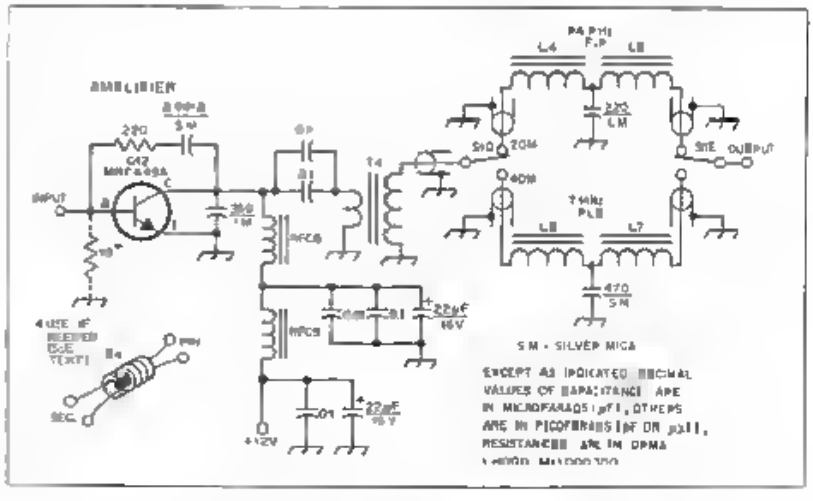
L6 — 12 turns no. 18 enam. wire on a T68-2 toroid core.

L7 — 14 turns no. 18 enam. wire on a T68-2 toroid core.

Q12 — Motorola MRF448A stud-mount transistor.

RFC8, RFC9 — 8 turns no. 18 enam. wire on an FT-50-43 ferrite toroid core.

T4 — Two rows of three each FT-50-43 toroid cores. Join with epoxy cement as shown in the inset drawing. Primary has one turn of no. 18 wire (U shaped). Secondary uses 3 turns of no. 18 enam. or insulated hookup wire. Primary leads exit from core at end opposite to secondary leads.



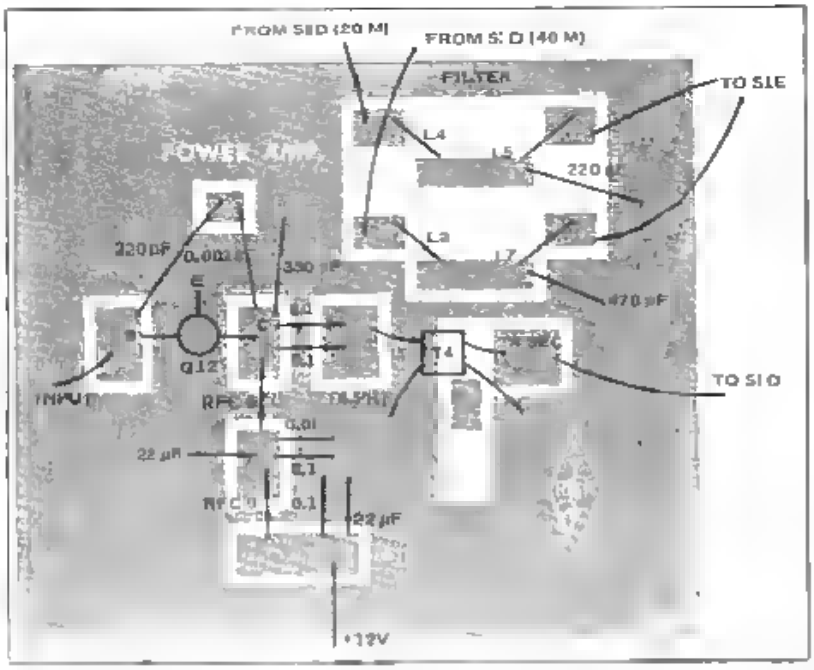


Fig. 11 — Parts placement guide for the PA pc board. Parts are mounted on the pattern side of the board, the shaded area in the view represents the copper pattern on the component side, and the other side of the board is unetched copper grounded plane. Decimal-value numbers alone represent capacitance in microfarads. Whole-number values with no units represent resistance in ohms.

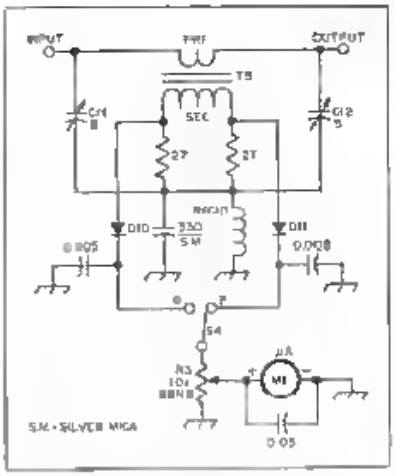


Fig. 12 — Circuit for the SWR sensor (see text). Fixed-value capacitors are disk ceramic. Fixed-value resistors are 1/2-W composition.
C11, C12 — 5-pF air variable, pc-board mount.
D10, D11 — 1N270 or 1N34A diode.
MI — Small microampere meter. A 50-, 100- or 200-µA type will be satisfactory.
R3 — 50-kΩ, linear-taper carbon control, panel mount.
RFC10 — Miniature 1-mH rf choke.
T5 — 50 turns no. 26 enam. wire on T50-2 toroid core. Primary has 2 turns of no. 26 enam. wire over center of secondary winding.

There are no special precautions to follow when assembling the amplifier, other than keeping the component leads as short as possible. Double-sided pc board should be used to minimize the chance for ground loops (feedback). They could cause amplifier instability.

The strip leads of Q12 should not be stressed when they are soldered in place. Allow a *slight* amount of slack for expansion when the transistor is heated during operation. Also, use care when tightening the transistor mounting nut. It should be drawn up just a "smidge" beyond the finger-tight point. A coating of silicone grease (heat-sink compound) should be placed on the transistor stud and metal face near the base of the stud. This will improve the transfer of heat between the heat sink and Q12. The heat sink is a homemade unit which has been bent into a U shape. It is made from a piece of 1/16-inch (1.6-mm) thick aluminum plate, 2-1/2 × 3 inches (64 × 76 mm) in size. Each lip is 1/2 inch (12.7 mm) high. The heat sink is affixed to the rear wall of the transmitter cabinet, and silicone grease is applied to the joining surfaces. The stud of Q12 and two no. 4-40 screws hold the heat sink firmly in place. The mounting method also holds the PA module in place on the inner surface of the rear wall of the cabinet. The pc-board layout is shown in Fig. 11.

SWR Indicator

As a convenience gadget we have included the SWR bridge shown in Fig. 12. It not only enables the operator to adjust the antennas for a low SWR when using a Transmatch, but serves as a relative-power-output indicator when switched to the forward mode. A blow-by-blow circuit description will not be given here, as this design was treated earlier in *QST* ("A QRP Man's RF Power Meter," June, 1973).

Assembly Notes — Composite Transmitter

Double-sided pc-board material is used for the cabinet of the WA6UZO version of the transmitter. Aluminum sheeting was bent into a U shape to form the W1FB prototype. The latter (HWD) is 3-1/4 × 5-3/4 × 6 inches (83 × 146 × 152 mm). The cover is a U-shaped piece of perforated aluminum. Two metal L brackets are affixed on the lower surface of the main chassis to permit the box cover to be secured by means of no. 6 sheet metal screws. The WA6UZO model of the transmitter is slightly larger than the W1FB version. He allowed room for mounting the modules horizontally. The vertical-mounting format makes it possible to realize greater miniaturization.

Our VFO is contained in a separate compartment. The enclosure is made from pc-board stock with the walls joined by means of solder. A U-shaped aluminum top cover is placed on the VFO assembly to prevent unwanted rf energy, moisture and dirt from entering. The

Rear view of the W1FB version of the transmitter. The heat sink is at the far right. The phone jacks and key jack are grouped at the left.

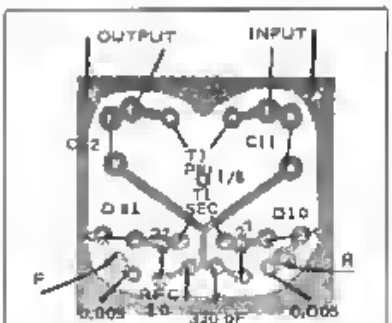


Fig. 13 — Parts placement guide for the SWR sensor. K = the cathode end of a diode. Fractional markings such as "1/8" identify mounting holes to be drilled with that size bit.

cover is press-fitted over the box walls. In the author's unit the SWR-sensor module is bolted to the VFO top cover.

By this time you should have a pretty good "handle" on how the collection of subassemblies are connected together. The block diagram of Fig. 1 provides the essential information. All of the signal leads should be made of shielded cable if they are more than two inches (51 mm) in length. RG-174/U subminiature coaxial cable is excellent for the purpose.

The main-tuning dial is a vernier mechanism (Calectro or Philmore). A large knob can be used as a substitute for the one which comes with the dial, as the original is a bit small for those who have large fingers. A Kurz-Kasch aluminum knob was used on the prototype model after its depth was reduced on a lathe (courtesy of W1SL).

The front and rear panels of the W1FB unit were sprayed a dark green color. Green Dymo tape labels were used to identify the controls. A reasonably professional appearance results from using labels which are the same color as the panel. Finally, four adhesive-backed plastic feet were affixed to the bottom of the cabinet.

Closing Remarks

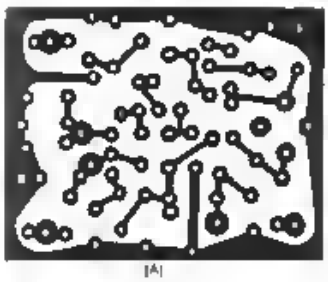
The toroid cores used in this project are available from Amidon Associates, G. R. Whitehouse and Palomar Engineers (check *QST* ads). It is suggested that the builder ask these suppliers for their catalogs, as some of the other components for the transmitter may be found in their product lines. It would also be prudent to scan the flea markets for parts.

The power supply for this transmitter should deliver 12 to 13 volts dc (regulated) at 3 amperes. Needless to say, a 12-volt car battery is suitable. A dry-battery pack is not recommended; the life span would be extremely short.

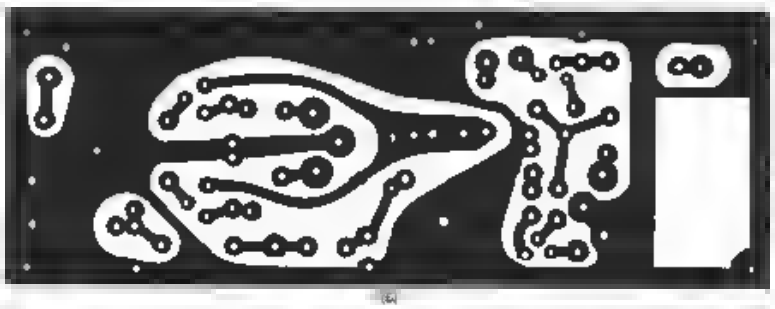
Motorola has included internal protection for their MRF449A transistor (Q12), so damage should not occur during short periods of operation when a mismatch greater than, say, 2:1 exists. This circuit has been tested into a dead short and a full-open load condition (key down) for periods of 30 seconds, and no damage to the PA stage resulted.

This two-band transmitter should provide many years of reliable operation. It is hoped that some useful information was passed along to those who aren't heavily immersed in solid-state design theory. If nothing more, let's hope we have negated the "Freddie syndrome" effectively.

Circuit-board etching patterns for the 7- and 14-MHz cw transmitter (DeMaw, "Transmitter Design—Emphasis on Anatomy" in four parts). Black represents copper. All patterns are shown at actual size from the foil side of the circuit board. See the drawings referenced below for parts layout information. At A, the VFO circuit board (Fig 3, Part 1). At B, the double-break-in delay board (Fig 5, Part 2). At C, the broadband amplifier board (Fig 8, Part 3). At D, the SWR sensor (Fig 13, Part 4).



(A)



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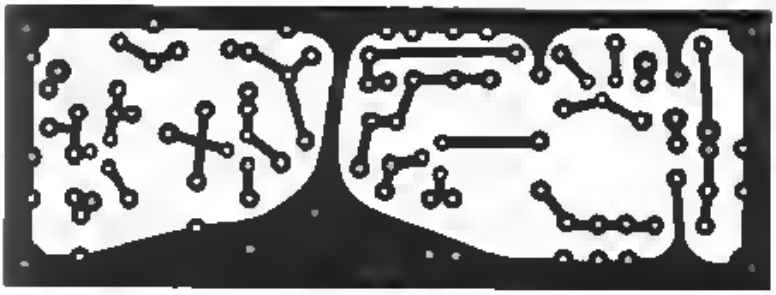

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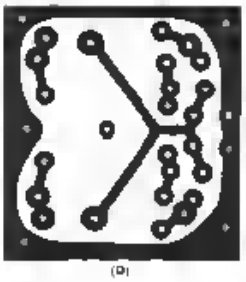
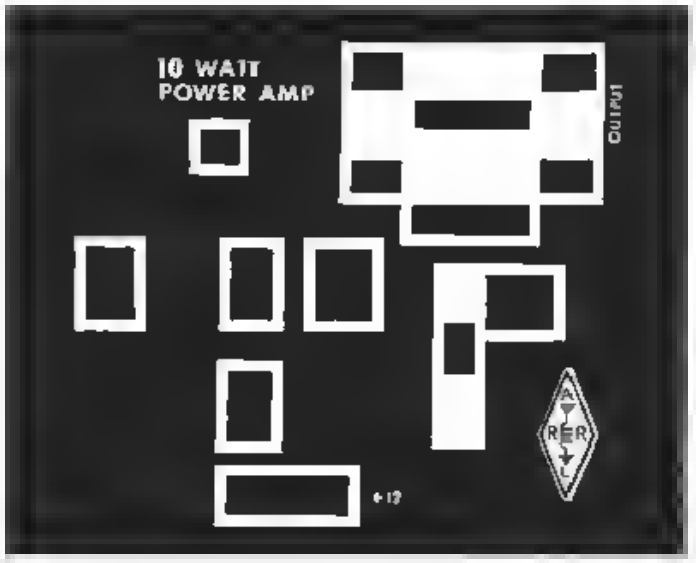


(D)

Circuit-board etching patterns for the 7- and 14-MHz cw transmitter (continued). Shown here is the power-amplifier circuit board, which appears in Fig 11, Part 4 of the series. This circuit board is double sided, the component-side foil being used only as a groundplane. That pattern is not shown as it contains only clearance holes for the component leads.



From February 1986 QST, p 23:

Four Watts, QSK, for 24.9 MHz

Here's your chance to try the 24.9-MHz WARC band at minimum cost. This transmitter is a fine mate for the 24.9-MHz converter described in April 1985 *QST*.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250, Luther, MI 49656

What might we expect from the new 24.9-MHz band? Well, it has similar propagation characteristics to the 10-meter band. It also exhibits some of the traits of the 15-meter band. Unfortunately, it is affected by sun-spot activity in a like manner to the other two bands above and below 24 MHz. Therefore, we are in a period of propagation ebb, owing to diminished sunspots.

Low power and reasonable antennas will do the job on 24.9 MHz as effectively as on 28 MHz. That is, it is not difficult to enjoy worldwide communications with less than 10 watts. With this thought in mind, plus an affinity toward being miserly when building a new rig, I designed the transmitter described here. You may build a duplicate model from scratch, or you have the option of purchasing a complete kit from a vendor.[1]

Circuit Details

The transmitter of Fig 1 features full-break-in operation (QSK). Operation requires only a key or keyer, antenna, a 12- to 14-V, 800-mA (or greater) regulated power supply (or car battery) and you, the operator. There is a terminal to which the receiver antenna line connects (terminal C of Fig 1).

Although crystal control is specified, a VFO can be substituted for Y1. Q1 is operated as a third-overtone oscillator. T1 and T2 are shielded transformers with tuned primary windings. They are arranged to provide an impedance transformation between the collectors and bases of the related transistors. This helps to ensure maximum RF-power transfer. The tuned transformers reject most of the unwanted harmonic energy before it reaches the driver and PA stages. You may substitute toroids[1] transformers and trimmer capacitors at T1 and T2 if you so desire.

Q3 serves as a broadband, class-A linear amplifier. It is the driver for the MRF475 power amplifier, Q4, which operates class C for maximum efficiency. A 7-section low-pass filter (FL1) is used as the output network to attenuate harmonic energy. The constants for FL1 were taken from *The ARRL Handbook* (see filter tables in the transmitting chapter). The power output from this transmitter is 4 watts into a 50-ohm load with an operating voltage of 12, and key-down current of 800 mA.

TR Switching

Q5 and Q6 of Fig 1 provide dc switching that enables the circuit to be classified as QSK. Q5 is a PNP keying switch that operates Q1 and Q2 for CW use. When the key is closed, Q5 triggers NPN transistor Q6 into the ON state, thereby shorting the receiver-antenna line (C) to ground during the transmit period. This prevents damage to the front end of the receiver or converter used with the transmitter. A similar technique was used by Wes Hayward W7ZOI to provide QSK operation. He used two reverse-connected 1N914 diodes as the shorting element during transmit. The measured RMS RF voltage on the receive-antenna line (key down) is approximately 0.4 with a 50-ohm termination. If diodes are used instead of Q6, the RMS voltage will be on the order of 0.7, key down.

C14 and L4 have a reactance of roughly 400 ohms. They serve as a series-tuned circuit to minimize loss of signal to the receiver during the receive period. FL1 serves as a filter ahead of the receiver, since the station antenna is attached to the output of FL1. Some insertion loss is present, but attenuation of the received signals is not significant.

S1 can be added to allow zero beating. It removes operating voltage from Q3, which helps lessen receiver overloading when you want to spot your transmitter signal. D2 is used as a dc gate to prevent the +12 V from reaching Q5, Q6 and the accessory terminal (J). The diode allows current to flow from Q5 to Q1 and Q2 (key down), but blocks the flow of current when S1 is set for the SPOT function.

S2 can be added for tune-up or Transmatch adjustments. If your key or keyer has a HOLD function, you may eliminate S2.

Key-down dc voltages have been noted at various points in the circuit of Fig 1. These have been added to aid in troubleshooting. The measurements were made with a Simpson 260 VOM. A 1-mH

Notes appear at end of article.

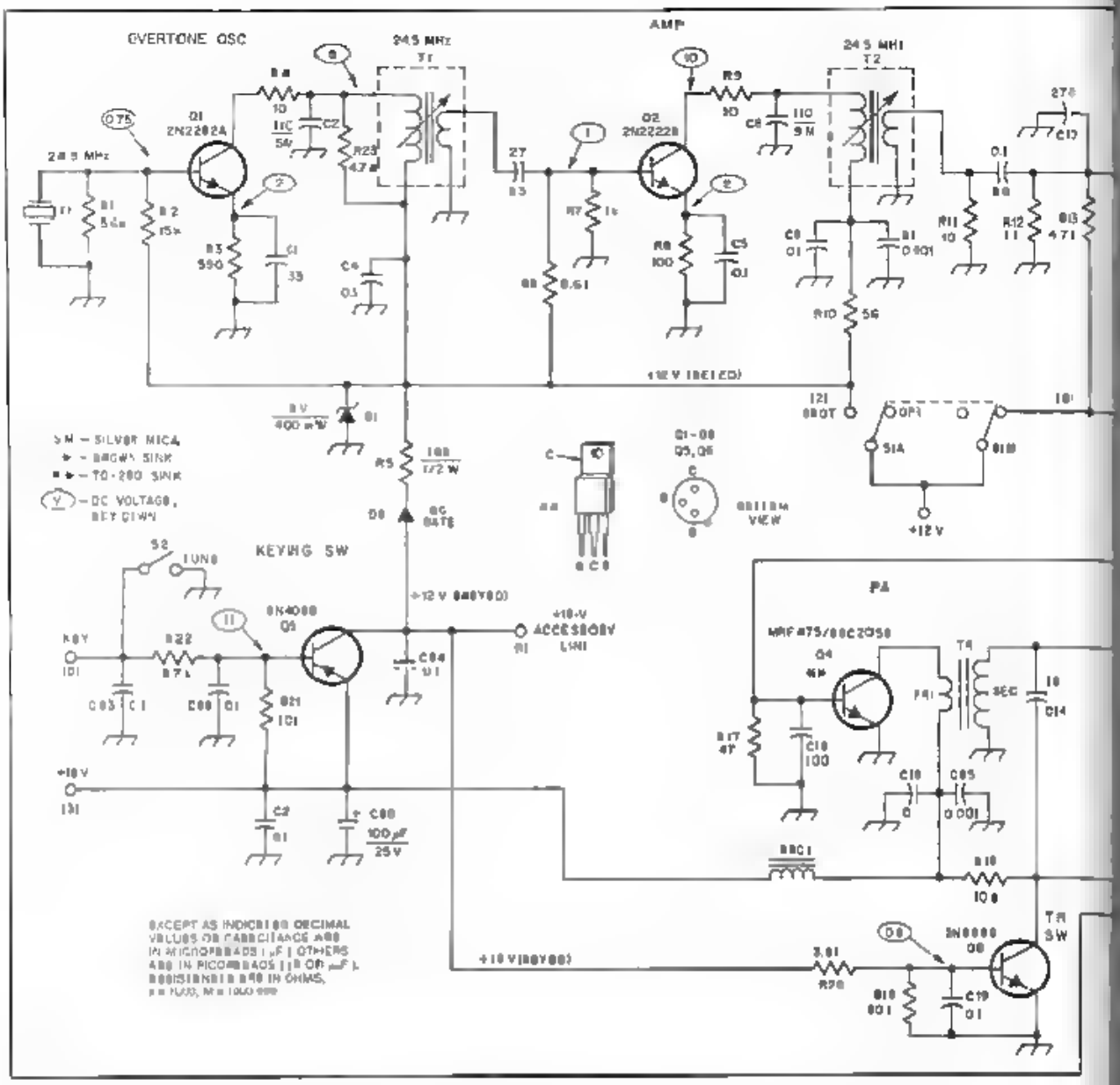
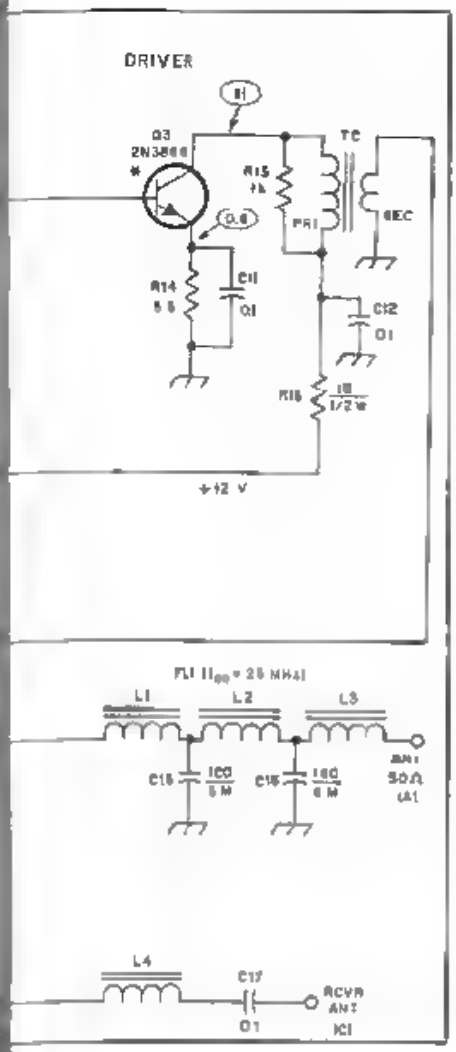


Fig 1—Schematic diagram of the 24.9-MHz transmitter. Capacitors are disc ceramic unless otherwise noted. Polarized capacitors are tantalum or electrolytic type. Resistors are ¼- or ½-W carbon composition units. Numbers inside circles indicate key-down dc voltages. Numbered components not appearing below are identified numerically for PC-board layout purposes only.

D1—6-V, 400-mW Zener diode.
D2—50 PRV, 1 A.
L1, L3—0.286-μH inductor. Use 8 turns of no. 24 enam. wire on an Amidon Assoc. T50-6 toroid core.
L2—0.5-μH inductor. Use 13 turns of no. 24 enam. wire on a T50-6 toroid core.
L4—L4 and C14 have reactances of 402 ohms. L4 is a 2.27-μH inductor. Use 24 turns of 26 enam. wire on T50-6 toroid core.
RFCI—Use 6 turns of no. 22 enam. wire on an Amidon Assoc. FT-37-43 ferrite toroid (two) FT-50-43 ferrite toroid cores. Use 10 turns of same wire for secondary.
T1, T2—Primary inductance is 8.38 μH. Use 6 turns of no. 26 enam. wire on bobbin of Amidon Assoc. L57-6 shielded transformer coil. Secondary has 4 turns of same wire.
T3—Broadband transformer. Primary contains 10 turns of no. 24 enam. wire on an FT-50-43 ferrite toroid. Use 2 turns of same wire for secondary. Spread secondary over all of primary.
T4—Broadband transformer. Primary has 7 turns of no. 24 enam. wire on stacked
Y1—Overtone crystal, 30-pF load capacitance, HC-6/U holder. International Crystal Mfg. Co. type G9, and plastic PC board mount holder. Choose frequency for lowest portion of the band. Do not attempt 12-m operation with a 12.450-MHz fundamental crystal. Substantial fundamental energy will appear in the transmitter output if Q1 is used as an oscillator/doubler.

RF choke was used between the positive lead of the VOM and the test point measured. This prevents unwanted RF energy from reaching the instrument and causing false readings. These voltages may vary slightly in accordance with the beta of the transistors used in your circuit. The RMS output voltage measured from (A) to ground across 50 ohms was 14. This indicates about 4 W of output power. Operating voltage was 12. My RMS measurements were made with a Hewlett-



Packard VTVM and RF probe that is rated to 900 MHz. However, an ordinary VTVM and homemade RF probe (see *The ARRL Handbook*) will work equally well.

Regulated voltage is ensured for Q1 by the addition of Zener diode D1; it sets the voltage level at +8. The lower oscillator-operating voltage helps to ensure frequency stability of Y1 by limiting the crystal current.

Parasitic suppression is aided by using R4, R9, C10 and C13. These components act as low reactances at VHF, but have little effect on circuit operation at 24.9 MHz.

Checkout and Operation

Our first assignment after completing the assembly is to give the PC board (non-component side) a thorough visual inspection to make certain we have no unsoldered joints or unwanted circuit bridges between unrelated PC-board foils. A magnifying glass is ideal for this step in the checkout. Make certain that all transistors are mounted correctly on the circuit board. Fig 2 indicates the placement of the transistor case tabs when the transistors are viewed from their tops.

Connect your power supply to the rig. Attach a 50-ohm load to (A). Place S1 in the SPOT position and switch S2 to OFF. Tune your receiver to the transmitter frequency. If a signal is heard, adjust T1 and T2 for maximum S-meter deflection. The transformer tuning will be broad, so don't be alarmed if the change in meter reading is small.

Place S1 in the OPR position and close S2 (TUNE). Measure the power output by means of an RF power meter, VTVM and RF probe, or oscilloscope with a 30-MHz or greater bandwidth.

If all systems are "go," key the transmitter and listen to the note in your receiver. The keying should be chirpless. If chirp is heard, adjust T1 for minimum chirp. Should this not resolve the problem, experiment with the value of feedback capacitor C1 until a clean CW note is heard. I tried three available crystals at Y1, and in all instances a good CW note resulted.

I purposely made the CW shaping a bit "hard." I have found this useful when operating at QRP levels. The shaping may be "softened" by changing the value of C23 (Fig 1). Start with a value of 1 µF. This will round off the trailing edge of the wave form. Increasing the capacitance of C24 will also affect the shaping.

Summary Remarks

The Motorola MRF475 may be difficult to locate. Other transistors of the same general specifications may be used at Q4. A 2SC2092 works well as a direct substitute and is available by mail.[2]

A scale template for the double-sided PC board is provided in Fig 3, parts placement is indicated in Fig 2.

There is no reason why this general circuit can't be modified for other amateur bands in the HF spectrum. All that needs to be changed are C1, the collector tuned circuits of Q1 and Q2, the constants of FL1 (see *The ARRL Handbook*), C14 and L4. Of course, Y1 must be chosen for the desired operating frequency.

VFO design data are contained in *The ARRL Handbook* and the ARRL book, *Solid State Design for the Radio Amateur* (out of print). I suggest that the VFO be operated at half frequency (12.45 MHz) to reduce the potential of chirp when the transmitter is keyed. A doubler stage (preferably a push-pull doubler) should be used to raise the VFO output frequency to

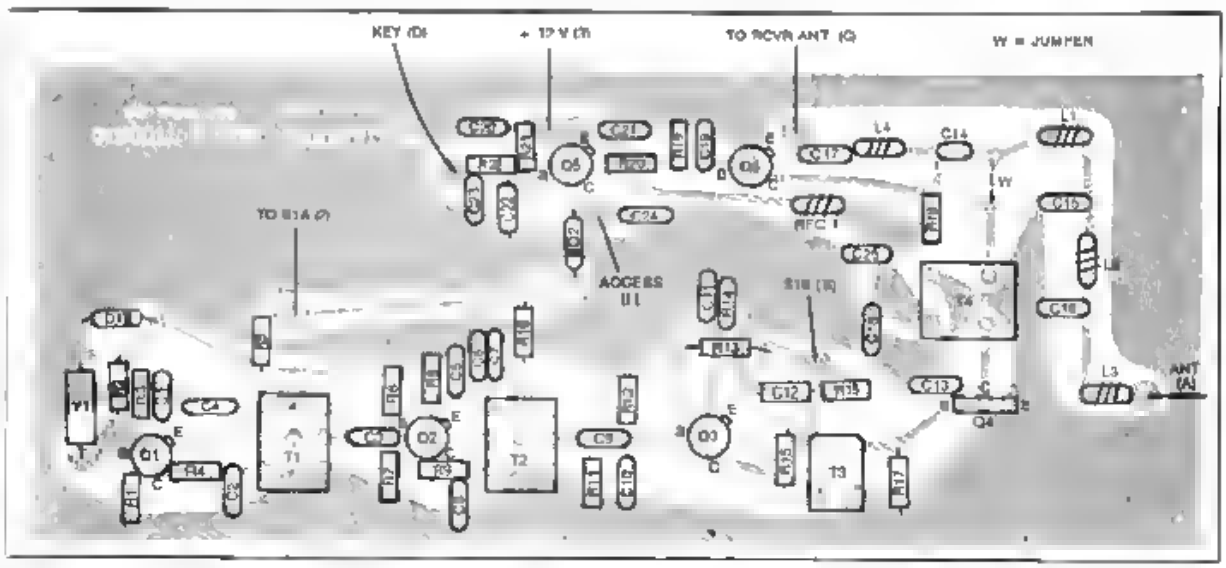


Fig 2—Parts placement guide for the 24-MHz transmitter PC board, as viewed from the component side of the board. R23 is mounted below the board on the appropriate solder pads.

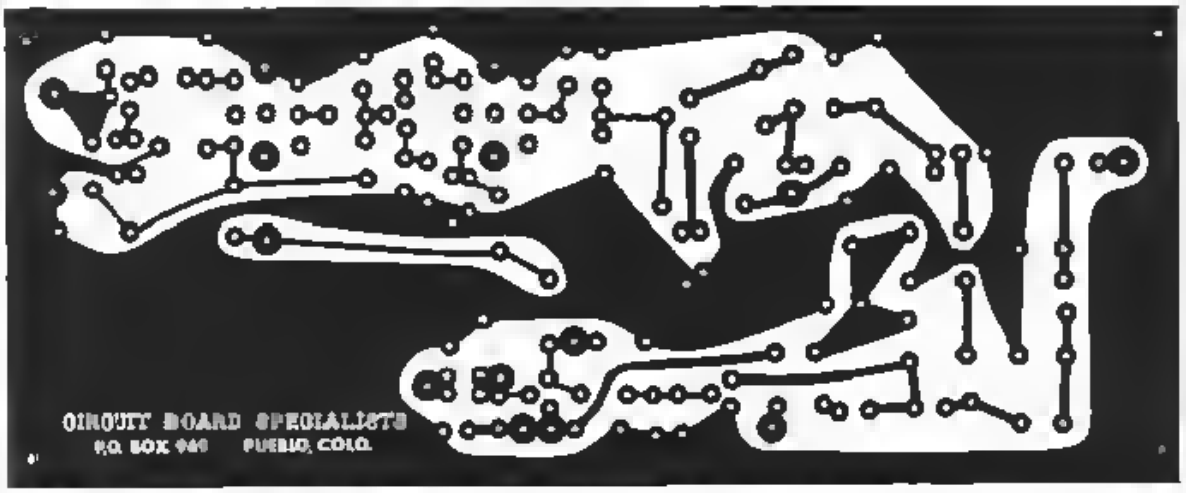


Fig 3—Scale template of the transmitter PC board as viewed from the etched side. Board is double sided, with the foil on the component serving as a ground plane. Connect etched ground foils to ground plane conductor at several points by means of component pigtails that pass through the board. Solder pigtails on both sides of the PC board.

1" (25.4 mm)

the 24-MHz band. VFO output should be approximately 2- to 3-V RMS across a 500-ohm load.

Even during mediocre propagation conditions I have found 24 MHz to be an interesting band. During 1984, I made numerous tests on the band with others while using an experimental license (KM2XQV) granted by the FCC. Many more QSOs were made under my amateur call after the 12-m band became available to us on June 21, 1985. Certainly, under skip conditions you should have a lot of fun with this little 4-watter! I hope to meet you on the new band.

Notes

[1]Chuck Hood, Circuit Board Specialists, PO Box 969, Pueblo, CO 81002, tel 303-542-5083. PC boards or complete kits available.

[2]Deleted.

For updated supplier addresses, see ARRL Parts Supplier List in Chapter 2.

From February 1988 QST, p 30:

Some QRP-Transmitter Design Tips

Full QSK is beneficial during QRP CW work. It is easy to achieve without relays at low power levels.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther, MI 49656

You may discover that full break-in (QSK) is an advantage for your QRP operating. It provides an opportunity to listen to your operating frequency during key-up periods. This lets you know if QRM is present, or if the other station is transmitting because the operator thought you stood by. (There may be times when your signals fade to such low levels that the person with whom you are communicating thinks you're standing by.) QSK can save wasted words in this situation. Full break-in is also beneficial during QRP Field Day operation. It saves time and can lead to a higher score.

This article is directed at those of you who like to build simple rigs. There is no practical project included, but the circuit in Fig 1 is a practical one. I built and tested the transmitter for the purpose of optimizing the performance, and to ensure that each stage operates as stated in this presentation.

Circuit Features

I will discuss the highlights of the Fig 1 circuit so you can understand how they work. This should help you design QRP transmitters on your own. Understanding the circuit functions is also useful when troubleshooting is necessary.

Refer to Fig 1. A VXO (variable crystal oscillator) is used at Q1 to generate the signal. Unlike most VXOs, this one takes the form of the familiar Pierce oscillator. I find this circuit more suitable for my needs than is the more common Colpitts VXO. The advantages are that no tuned output circuit is required to develop adequate excitation for the subsequent RF stage. Also, C2 (frequency control) will swing the crystal frequency *above and below* the marked value. Most Colpitts VXOs do not allow the crystal to be "rubbered" above the marked frequency. My tests were made with an AT-cut plated crystal in an HC-6 holder (International Crystal Mfg Co no. 433113) with a marked frequency of 7050 kHz. The load capacitance of Y1 is 30 pF. C2 of Fig 1 permits the crystal frequency to be moved from 7045 to 7052 kHz. Greater inductance at RFC1 will allow a wider frequency shift, but at the cost of frequency stability. The 7-kHz swing yields crystal controlled stability, even during wide excursions of ambient temperature. This is important when operating QRP during Field Day or on camping trips; vast temperature changes may occur from day to night. The negative feature of the Fig 1 VXO is that C2 must be insulated from ground. In other words, both the rotor and stator must be above ground. The tuning capacitor can be mounted on a plastic bracket to achieve isolation.

R5 and R13 of Fig 1 are used to lower the Q of RFC1 and RFC2. Too great a Q causes crud to appear at the leading edge of the keyed waveform (spurs). The resistors cure this problem. C1 is a feedback capacitor. The value is chosen to provide chirpless keying and high output from Q1. You may need to experiment with the C1 value. The crystal activity and the gain of your particular Q1 transistor will dictate the optimum value for C1.

RF Power Amplifier

I like to experiment with transistors that are not intended to be used for RF applications. The Motorola MPS-U02 is an example, a device that was designed for audio and switching use. It is frequently used as one half of a complementary symmetry audio amplifier (paired with an MPS-U52). The f_T (upper frequency limit) is 150 MHz, and it can handle up to 800 mA of continuous collector current. The specifications strongly suggest RF power use! The maximum V_{ceo} (collector to emitter voltage, base open) is +40. This allows plenty of leeway for the collector voltage to swing beyond 12 volts in RF or audio service. Typically, the RF collector voltage (sine wave) will rise to twice the power supply value, or 24 volts for a 12-V dc supply during CW operation.

The cost for MPS-U02s is quite low—another advantage. I bought 10 of them as surplus for 39 cents each. They are listed as new devices (89 cents each) in the Circuit Specialists catalog.[1] Numerous other high f_T audio/switching transistors are suitable for RF power amplifier use as well. Pick a device that has an f_T of five or more times the operating frequency. This will ensure ample gain at the desired frequency.

I used simple capacitive coupling between Q1 and Q2. C4 is selected to provide 1.5 watts of output from Q2 for my circuit. I needed 33 pF of capacitance. Larger values will increase the transmitter power, but at the risk of exceeding the safe ratings of Q2. The light coupling provided by C4 minimizes oscillator loading. Too great a value at C4 can kill the oscillation of Q1. I chose the 1.5-W output power to cause the Q2 collector impedance to be 48 Ω. This is determined from $Z = V_{cc}^2/2P_O$, where V_{cc} is the collector to emitter dc voltage, and P_O is the power output. This enabled me to use a 50-Ω filter (FL1) without a broadband matching transformer between Q2 and FL1. A heat sink is required on the tab of Q2 to minimize the transistor junction temperature. A 1-inch

[1]Circuit Specialists, PO Box 3047, Scottsdale, AZ 85257. Phone 1-800-528-1417 when ordering. Catalog available.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

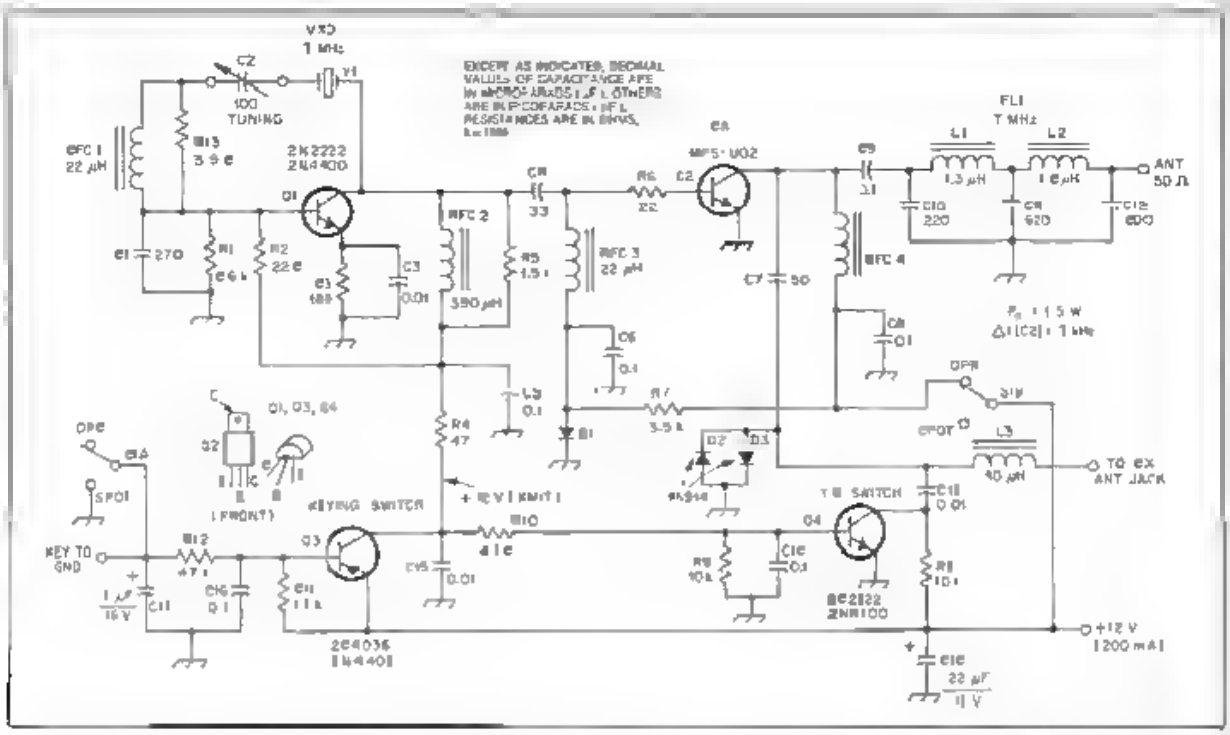


Fig 1—Schematic diagram of the test transmitter. Fixed-value capacitors are disc ceramic unless otherwise noted. Polarized capacitors are tantalum or electrolytic. Resistors are ½-W carbon composition

C2—100-pF miniature air variable or 10-100 pF compression trimmer with shaft
C10, C11, C12—Polystyrene or silver mica.
D1—Rectifier diode, 50 PRV, 1 A
D2, D3—Small-signal switching diode, 1N914 or equiv
L1, L2—1.3-µH inductor. Use 18 turns of no. 26 enam wire on an Amidon Assoc T-37-2 toroid
L3—10-µH inductor. Use 45 turns of no. 28 enam wire on an Amidon Assoc T-50-2 toroid
RFC1, RFC2, RFC3—Miniature ferrite core RF choke.
RFC4—12 turns of no. 26 enam wire on an Amidon Assoc FT-37-43 ferrite toroid (650 µ).
S1—DPDT toggle or slide switch.
Y1—Fundamental crystal (see text).

square piece of no. 16 gauge aluminum or copper was suitable for my test circuit. A 10-minute key-down period at 1.5 W output caused the transistor and heat sink to be moderately warm to the touch.

Harmonic Filter

The FCC permits emissions requirement is more liberal at power outputs under 5 W. A five-section low-pass filter (FL1 of Fig 1) is ample to comply with regulations. A seven-element filter would offer greater attenuation of the 2nd and 3rd harmonics, should that be your desire. My filter constants were obtained from the normalized filter tables in the transmitting chapter of *The ARRL Handbook*. I chose an f_{co} (cutoff frequency) of 8 MHz. The ripple factor is 0.01 for FL1. The ripple indicates the relative flatness (lack of amplitude dips and peaks) of the peak portion of the filter response curve.

If FL1 were terminated at each end with a 50-Ω nonreactive (purely resistive) load, C10 and C12 would have the same value. However, we must recognize the transistor output capacitance (20 pF for an MPS-U02), stray circuit capacitance (roughly 10 pF) and the value of the TR sampling capacitor, C7. During key-down periods, C7 is switched in parallel with C10 via D2, D3 and Q4. The approximate total of these capacitances is 80 pF. This value must be deducted from 300 pF (normal C10 and C12 value) if the filter is to perform properly. A 220-pF capacitor is, therefore, appropriate for C10. Additional harmonic reduction results from operating Q2 in the class-A linear mode. D1 provides approximately 0.7 V of forward bias for Q2. The class-A mode reduces the excitation requirement for Q2, which is also a benefit. A similar RF amplifier, operated in class C, would require significantly more RF drive, and the harmonic output would be somewhat higher in amplitude.

TR Circuit

You will note in Fig 1 that a TR (transmit-receive) circuit is included. Q3 is a PNP dc switch that applies operating voltage to Q1 when the key is closed. Key closure shorts the base of Q3 to ground, and this causes it to conduct. During conduction, +12 V is connected to Q1 through the junction of Q3. The keyed +12 V is routed also to NPN switch Q4. This transistor also conducts when the key is closed. At full saturation the Q4 collector-emitter junction closes and thus shorts the receive antenna line to ground through Q4. D2 and D3 also accomplish this function, but leave a residual RF voltage of 0.7 V RMS on the receive line. The shunt-diode technique was popularized by Wes Hayward, W7ZOI, in some of his QRP transmitters that featured full QSK. I use the diodes as backup protection, should Q4 fail to operate for some reason.

TR circuit sampling capacitor C7 should have a reactance no less than 400 Ω. Smaller reactance values will rob transmitter output power when the key is closed. Some power is sacrificed with the value shown for C7, but it is minimal. The trade-off associated with this type of TR circuit is a slight signal loss during receive, owing to the small value for C7. Both Hayward and Lewallen (W7EL) reduced this problem by adding L3 in the receive antenna line. L3 has the same reactance as C7. This permits C7 and L3 to form a series-resonant circuit at the operating frequency, which in turn reduces the loss in the receive signal that is fed to the receiver. A slug-tuned coil (variable inductor) at L3 would help to make the series circuit exactly resonant.

I measured the RF voltage from the receive antenna line to ground with a Tektronix 453A scope during key-down. It is 200 mV P-P (70.7 mV RMS) across 50 Ω. This potential will not harm any receiver, solid-state or tube type.

Additional TR control is possible if you connect an outboard NPN switch to the keyed +12 V (between Q3 and Q4). The

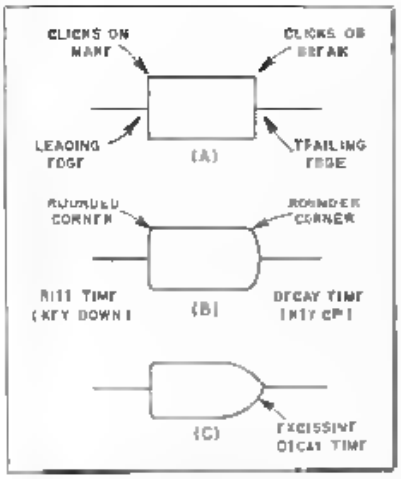


Fig 2—Examples of keyed RF waveforms. The illustration at A shows hard keying with square waveform corners. This waveform causes clicks on the make and break of the key. An acceptable waveform is shown at B. The corners are rounded to remove clicks and the decay time is lengthened somewhat over that shown at A. Soft keying is shown at C. The decay time has been increased over that at A and B. This waveform is not suitable for high speed keying. See the text for additional data.

outboard transistor switch can then be used for receiver muting, or for actuating a small 12-V relay, which may, in turn, serve as a receiver muting control.

The Keyed Waveform

Many homemade QRP transmitters are deficient in harmonic suppression and keyed wave shaping. I have been lax in the latter regard myself. One tends to justify hard keying as being somewhat more effective at the QRP level, and in a sense this is true. However, under no circumstances should the keyed wave cause clicks. The dividing line between acceptable hard keying and clicky keying is rather thin! It is better to stay on the safe side and attempt to obtain a keyed wave that has a 5 ms rise and fall time, which is considered entirely acceptable. This represents a clickless wave that has a fast enough recovery time to permit very high speed keying. Too long a decay time (key up) will limit the useful keying speed we can apply. Fig 2 shows a hard, clicky wave at A. The drawing at B illustrates a wave with rounded corners that does not cause clicks. Fig 2C shows a soft wave with a long tail. This waveform is unsuitable for fast keying.

The waveform from your transmitter can be examined by sampling the transmitter RF output energy across a 50-Ω load. Rapid keying of the transmitter will cause the RF envelope to be displayed on the scope tube.

Waveform shaping is accomplished in Fig 1 by means of C16, C17, R11 and R12. Bypass capacitors in the keyed circuit (such as C5 and C15) also affect the shaping.

The decay time (trailing edge of the waveform) is affected by C16, C17, R11 and the bypass capacitors mentioned above. R12 affects the attack time (leading edge of the waveform). In fact, you may add additional resistance between R12 and the key jack to shape the leading edge of the waveform. Values up to 10 kΩ are suitable. The larger the resistance of R11, the slower the waveform decay time. The R11 value shown allows the base of Q3 to return quickly to +12 V, thereby cutting off the Q3 conduction (key up) quickly. This fact was brought to my attention by Ed Hare, KA1CV, of the ARRL lab staff.

The shaping-network values in Fig 1 ensure a keyed waveform that is clickless, but hard enough to give "presence" to the CW note. The frequency-control values for the VXO in Fig 1 prevent the signal from sounding chirpy when the VXO is keyed by Q3.

Final Comments

I added S1 to facilitate frequency spotting without placing the transmitter on the air. S1A closes the key line to turn on Q3. S1B removes operating voltage from Q2 at the same time. This reduces the signal strength of the beat note heard in my receiver. In other words, it is not so strong that it overwhelms my receiver. S1B also prevents the transmitter signal from reaching the antenna during zero beating or spotting.

You may feel that a VXO is not nearly as desirable as a VFO. I confess that 7 kHz of frequency swing is a small amount, but the VXO is stable under most conditions, and this appeals to me during operation afield. It is not a severe handicap to carry two or three crystals when camping. This provides sufficient frequency coverage of the 40-meter band. In fact, you may wish to include a low-capacitance crystal selector switch if you build a VXO rig of this type. But remember that the more stray capacitance you introduce in the crystal circuit, the smaller will be the frequency swing of a given crystal.

My purpose in writing this article is to pass along some design hints that you may not have considered. The points I have covered are among the most frequently asked questions I receive concerning QRP transmitters. The main point I want to make is that you *can* build your own gear, and it takes little additional time or money to develop a circuit that operates cleanly and reliably.

From November 1986 QST, p 43:

A QRP Transmitter for 30 Meters

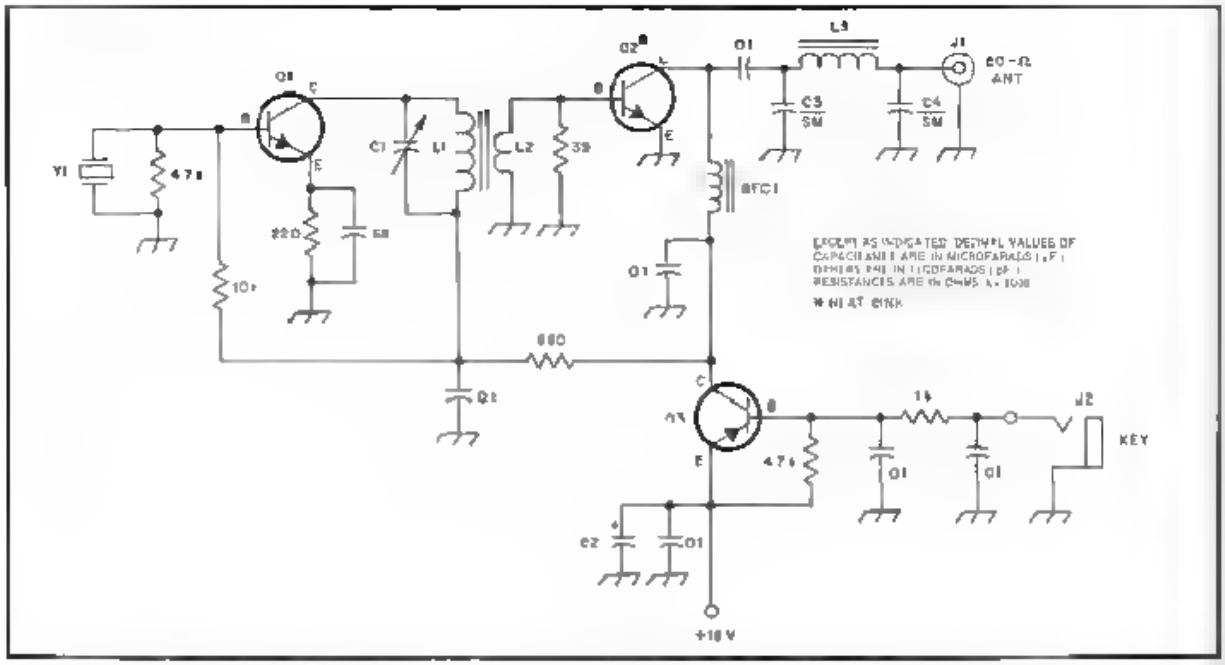


Fig 1—A schematic diagram of the QRP transmitter for 30-meter operation. Fixed-value capacitors are disc ceramic. Inductor cores are available from Amidon Associates or Palomar Engineers.[1,2] The enclosure is from Radio Shack[3] (RS-270-251), and the circuit board is from Circuit Board Specialists.[4] An etching pattern for this circuit appears on p 11.

Y1—Fundamental-mode crystal for the 30-meter band.
Q1—2N2222A or equiv.
Q2—2N3553 or equiv.
Q3—2N4036 or equiv.
C1—150-pF mica trimmer, ARCO no. 424.
C2—470-μF 25-V electrolytic or tantalum capacitor.
C3, C4—330-pF silver-mica or polystyrene capacitor.
L1—30 turns, AWG no. 24 enameled wire on a T-50-2 core.
L2—3 turns, AWG no. 24 enameled wire over L1.
L3—13 turns, AWG no. 22 enameled wire on a T-50-2 core.
RFC1—30 turns, AWG no. 28 enameled wire on an FT-37-62 core.
J1—SO-239.
J2—Phone jack or phono jack.

[1]Amidon Associates, 12033 Otsego St, North Hollywood, CA 91607, tel 213-760-4429.
[2]Palomar Engineers, PO Box 455, Escondido, CA 92025, tel 619-747-3343.
[3]Circuit Board Specialists, PO Box 969, Pueblo, CO 81002, tel 303-542-5083.
For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

☐ Fig 1 shows an inexpensive transmitter for the 30-meter band. The combination of excellent propagation characteristics and a relatively low QRM level on this band make solid communication routine at QRP levels. The circuit shown was adapted from a W7ZOI design shown in "Experimenting for the Beginner" by Doug DeMaw, W1FB, in the September 1981 issue of *QST*. The transmitter can be put on 30 meters with relative ease.

The only major changes required were to resonate the oscillator output circuit, and filter the amplifier output on the new frequency. This was done by changing the number of turns to L1 and selecting an appropriate range for C1. A new output filter (C3, C4, L3) was designed using the component values shown in the schematic. In addition, the antenna output jack was changed to an SO-239 (to suit my personal preference), and a ground strap was added to the enclosure (see Fig 2). My circuit is built on a printed circuit board as in the original article. The transistors shown differ from the original design only because they were available in my junk box.

The transmitter performance is excellent, and the keying is clean. Output power is exactly 1 watt when using a 12.6-V dc supply (measured with a VTVM and a 50-Ω resistor). Excellent signal reports have been received from stations in New Brunswick, Florida and from as far west as Kansas. —*Frank Pitmar, WD4DDS, 12 E Lakeshore Dr, Rome, GA 30161*

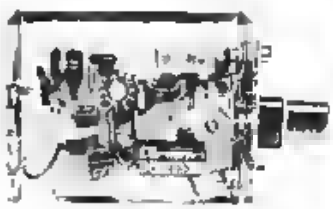


Fig 2—An interior view of the QRP transmitter as converted for 30 meters. The heat sink is on Q2.

From February 1984 QST, p 46:

A Two-Transistor Transmitter for 30 Meters

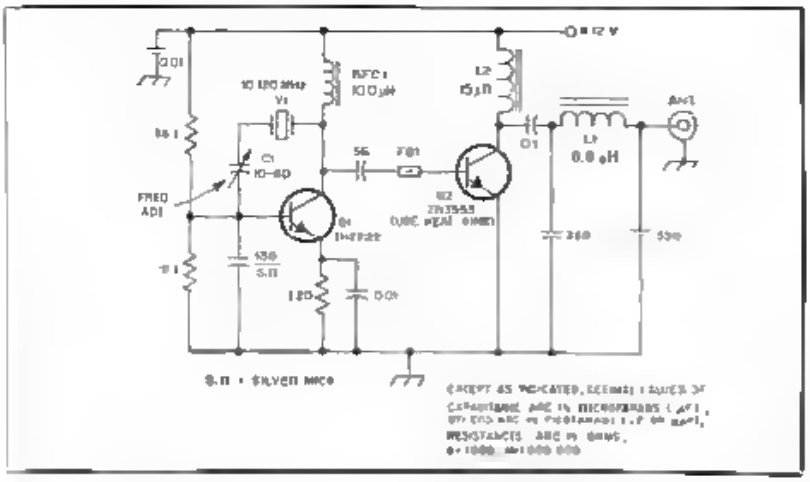


Fig 1 — Schematic diagram of a simple 30-m transmitter built by KB4PY. Resistors are ½-W, carbon-composition types. Capacitors are disc ceramic, except C1.

C1 — 10- to 30-pF mica trimmer.
FB1 — FB43-101 ferrite bead.
L1 — 13 turns no 22 enameled wire on a T68-6 toroid core (yellow). 0.8 µH.
L2 — 30 turns no 25 enameled wire on an FT37-63 toroid core, 15 µH.
RFC1 — 100-µH RF choke, such as Miller part no. 4632.
Y1 — Fundamental-type crystal in FT-243 holder for any frequency in the 10-MHz amateur band.

came up with the one shown at Fig. 1. The circuit is simple and inexpensive. It uses fundamental-type crystals in FT-243 holders, which are easy to obtain from a variety of sources. The transmitter output is almost 1.5 W, and the harmonics are 34 dB down from the fundamental. With a reasonably stocked junk box the total cost for the transmitter should be under $10.

My antenna system is a 150-ft random wire and a Transmatch.[2] A two-position coax switch is used to change between transmit and receive. I leave my receiver active during transmit, and it provides a nice sidetone. Keying either the positive or negative power supply leads seemed like the easiest method to me. The keying waveform is a bit soft, but I don't believe the slight chirp is objectionable. Others may wish to experiment with alternate keying methods.

I built my rig on a piece of perf board, and mounted the circuit inside of an aluminum box, as shown in Fig. 2. An etching pattern and parts-placement diagram are shown in Fig. 3. Many of the parts are available from RADIOKIT. Circuit Board Specialists also has a PC board and a complete kit of parts available for this project.

I used an oscilloscope and frequency counter to align the circuit. Alternatively, use a calibrated receiver, and tune C1 for maximum output. I hope others have as much fun with this little rig as I have.—*Paul Hoffman, KB4PY, 4502 Indian Hills Rd, Decatur, AL 35601.*

1m = ft × 0.3048, mm = in × 25.4

[2] When I decided to become active on the 30-m band, I wanted to build a simple transmitter. I have a Yaesu FT-101E that receives WWV on 9 MHz. Others may have general-coverage receivers, and need only a transmitter. April 1983 *QST* described an elaborate 30-m rig.[1] But even the transmitter section is more than just a "junk box" project.

After testing a variety of other circuits, I finally

[1] D. DeMaw, "Putting the 'SP6 Special' Transceiver Rig on 10 MHz," *QST*, April 1983, pp 19-21.

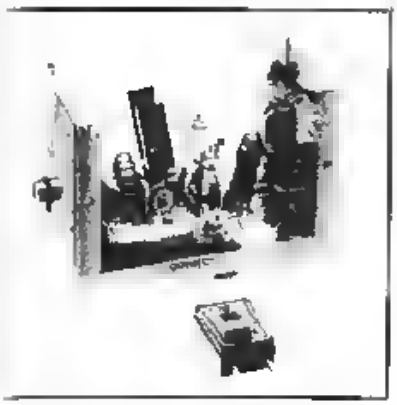


Fig 2 — Photo showing the construction technique used by KB4PY on his 30-m QRP transmitter. Note the heat sink used on Q2. If you use point-to-point wiring, as shown, it may be necessary to reduce the value of the 30-pF silver-mica capacitor. The author used a value of 82 pF in his original design.

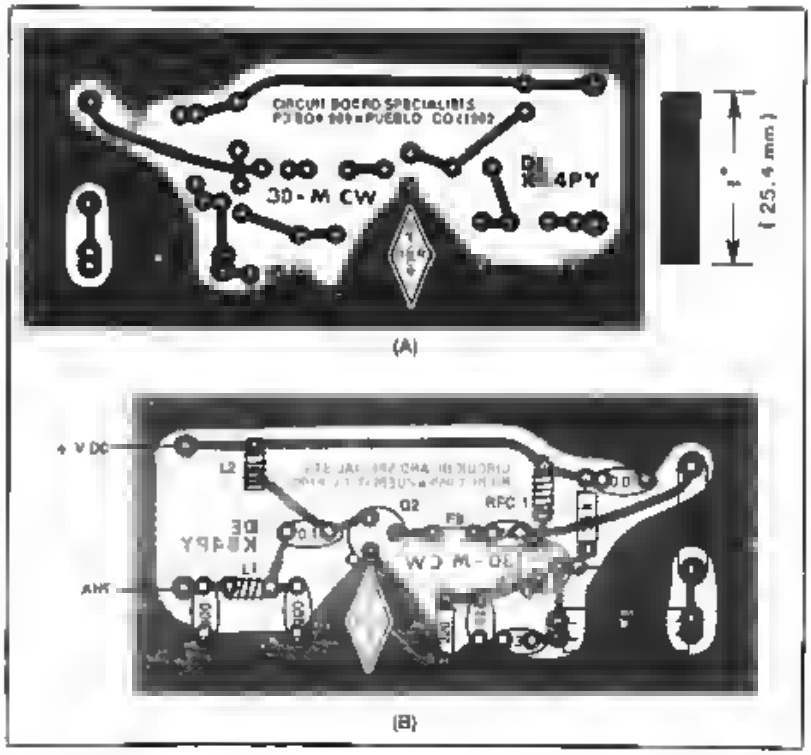


Fig 3 — A PC-board etching pattern is given at A. Black represents unetched copper, viewed from the foil side of the board. This pattern is shown actual size. B shows a parts placement diagram. Components are placed on the non-foil side of the board, the shaded area represents an X-ray view of the copper pattern.

From May, 1979 QST, p 27:

A VMOS FET Transmitter for 10-Meter CW

Vertical metal-oxide semiconductor FETs are new on the amateur scene. Here is a practical construction project that makes use of the Siliconix VN88AF.

By Wes Hayward,* W7ZOI

Although QRP cw operation has been the major passion at W7ZOI for many years, 10 meters is a band that has been bypassed. The reason is not clear, for it's hard to find a better frequency during periods of high sunspot activity. The rig described here is a long-overdue remedy for this neglect.

It was decided to try one of the new vertical metal-oxide semiconductor field-effect transistors as a power amplifier, rather than to use a conventional approach to transmitter design. Experiments with earlier VMOS FETs were encouraging. However, the devices were either expensive or completely unavailable. Today, plastic medium power devices are readily available for less than $2.

The transistor chosen was the Siliconix VN88AF. With 80-volt drain-to-gate and drain-to-source breakdown voltages and a peak current capability of 3 amperes, the device appeared ideal. The major limitations are the power dissipation of 15 watts and the presence of a protection Zener diode at the gate. The latter turned out to be a major constraint for cw operation. (We'll have more comments about that later.)

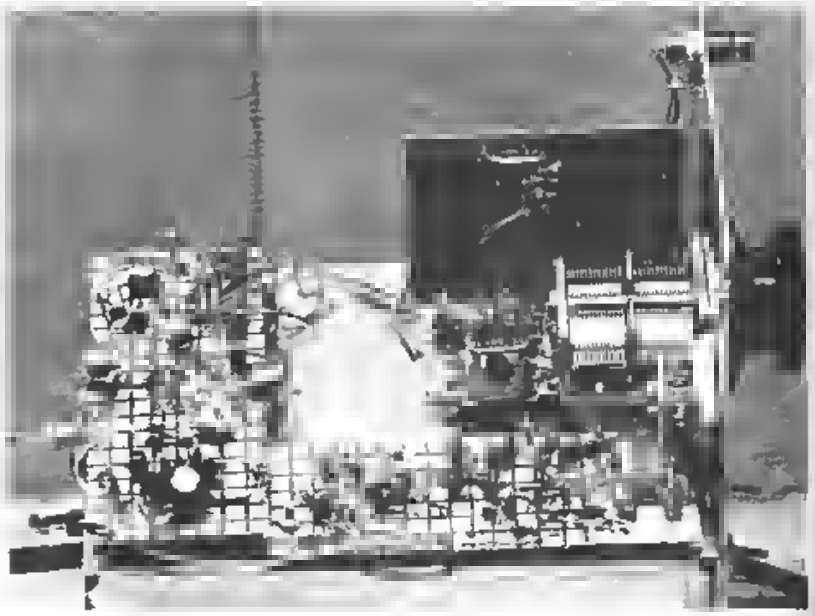
Modern operating practices dictate the need for some degree of frequency agility. A 14-MHz VXO was chosen for frequency control. A clean balanced doubler provides the required 28-MHz signal. Low-level stages with an abundance of stabilizing negative feedback increase the power to drive the VMOS final. All indications are that the system should be eminently reproducible.

Circuit Details

The heart of the transmitter, the rf chain, is shown in Fig. 1. Q1 serves as a crystal Colpitts oscillator with the crystal operating on the inductive side of resonance. The crystal normally used in this circuit has a marked frequency of 14,025 kHz. With the components shown, a 25-kHz range is obtained at 14 MHz. (Some experimentation may be required with the number of turns on L1 to obtain the desired range.) As shown, the circuit tunes from 14,025 down to about 14,001 kHz. If the inductor L1 is shorted, the circuit will tune from the marked crystal frequency upward about 10 kHz.

Not all crystals will function well in this

Side view of the 10-meter VMOS FET transmitter. No attempt was made to miniaturize the unit. The final amplifier, Q5, is visible at the lower left. The crystal is immediately to the left of the variable capacitor. The connector at the upper right is for the receiver.



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†Notes appear at end of article

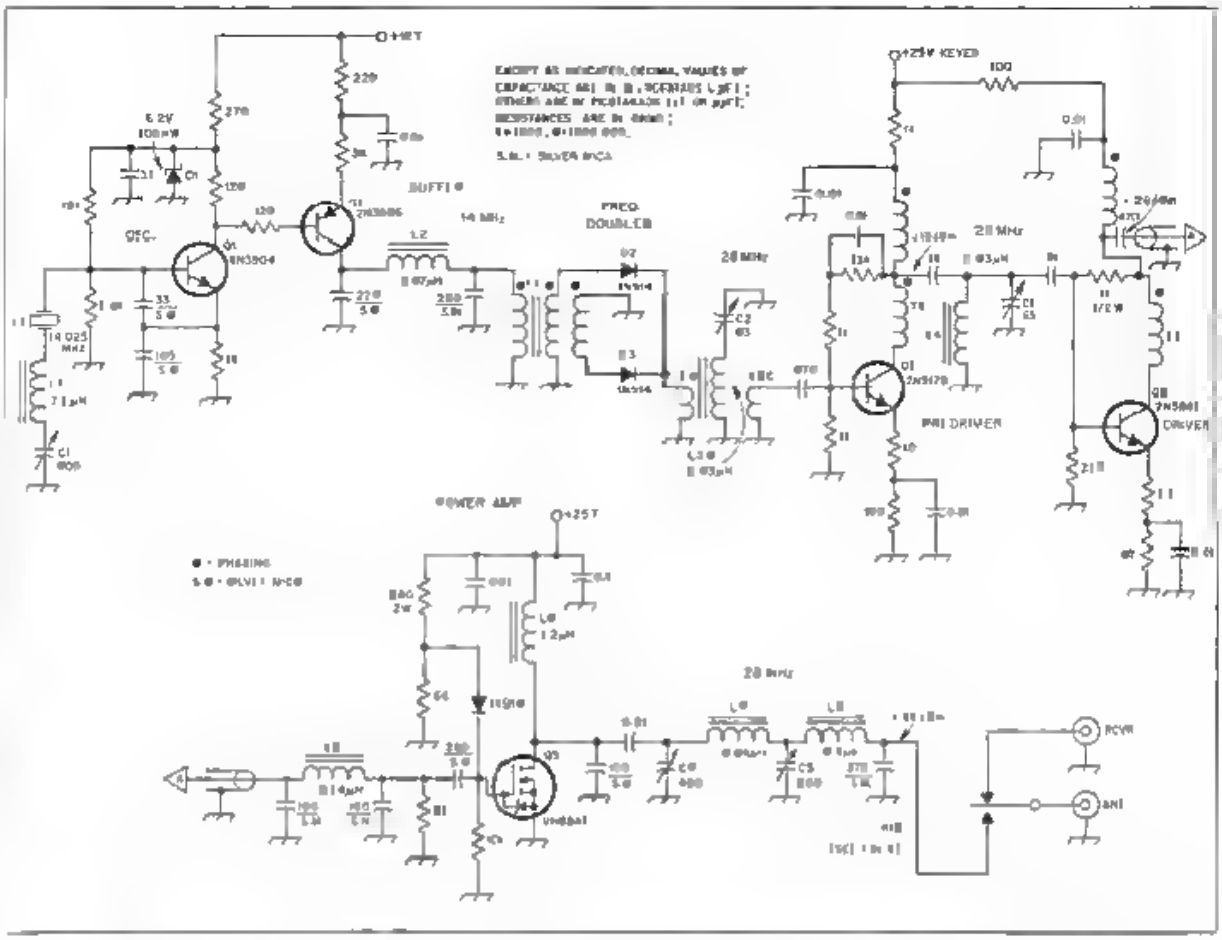

Fig 1 — RF chain for the VMOS FET transmitter. All resistors are ½-watt composition and capacitors are disk ceramic unless otherwise specified. Tolerances are ±10 percent. The +12 T indicates 12 V during transmit periods only (see Fig 2). The Supertex VN5-A is an equivalent part for the VN88AF (Siliconix) VMOS transistor. Siliconix parts are almost impossible to purchase in small quantities, but Hamilton-Avnet outlets seem to have a limited number of the transistors for single-lot purchases.

C1 — 400-pF air variable. Allied 895-4260 or equiv.
C2, C3 — 95 pF trimmer, Allied 782-0404 or equiv.
C4, C5 — 400 pF trimmer, two Allied 782-3765s in parallel or equiv.
D1 — 6.2 V, 400-mW Zener diode, 1N753A or equiv.
D2, D3 — 1N914 or equiv.
L1 — 7.2 µH, 40 turns no. 24 wire on Amidon T-80-6 core.
L2 — 0.87 µH, 12 turns no. 26 wire on Amidon T-37-6 core.
L3A, L4 — 0.43 µH, 12 turns no. 24 wire on Amidon T-37-6 core.
L3B, L3C — 2-turn links over L3A.
L5, L7 — 0.24 µH, 9 turns no. 24 wire on Amidon T-37-6 core.
L6 — 0.3 µH, 10 turns no. 24 wire on Amidon T-37-6 core.
L8 — RF choke, approx 1.3 µH, 17 turns no. 24 wire on Amidon FI-41-6 core.
T1 — 7 bifilar turns no. 28 wire on FT-37-43 ferrite core.
T2, T3 — 7 bifilar turns no. 28 wire on FT-37-43 ferrite core.
Y1 — See text.

circuit. They should be fundamental-mode units, the usual case at 14 MHz. The best results are obtained with HC-6/U metal-can units, such as those manufactured by JAN Crystals and by International Crystal (type 031300). The most reliable operation occurs when the metal crystal case is grounded; if it is allowed to "float," the frequency will change when a hand moves near the rock, making a front-panel mounted crystal socket impractical. Tuning is very nonlinear, but this presents no problem in this application. The 50-kHz tuning range (after doubling) has been more than sufficient. While a 400-pF variable capacitor is used, a smaller unit will suffice with only a slight reduction in tuning range. The power available from the oscillator is around one milliwatt (0 dBm).

Q2 functions as a buffer amplifier to increase the 14-MHz power to about +10 dBm, a near-optimum drive level for the diode doubler. The output of this stage has a low-pass filter to ensure a waveform relatively free of harmonics that would degrade the balance of the multiplier and hence reduce the suppression of 14-MHz energy in the output.

Frequency multiplication is obtained with a pair of silicon diodes, D2 and D3. One might question the use of a passive frequency doubler, but careful experiments using laboratory instrumentation have confirmed the wisdom of this choice. Details of this work are presented in chapter 3 of *Solid State Design for the Radio Amateur*.[1] The method is used in several projects in that book.[2]

The doubler is followed by a single tuned circuit at 28 MHz. A pair of two-turn links on the toroidal inductor couple energy into and out of this resonator. The power available from the doubler, after filtering in the resonator, is about 0 dBm.

The 28-MHz energy is applied to a two-stage, keyed amplifier. Negative feedback is used in both stages to ensure broadband stability and to establish the gain levels desired. The resistor values used were chosen from a program written for the writer's programmable calculator. Additional information on feedback amplifiers is presented in chapter 8 of *Solid State Design*. The saturated output of Q4 is nearly 1/2 watt, more than enough to drive the VMOS final amplifier. Both

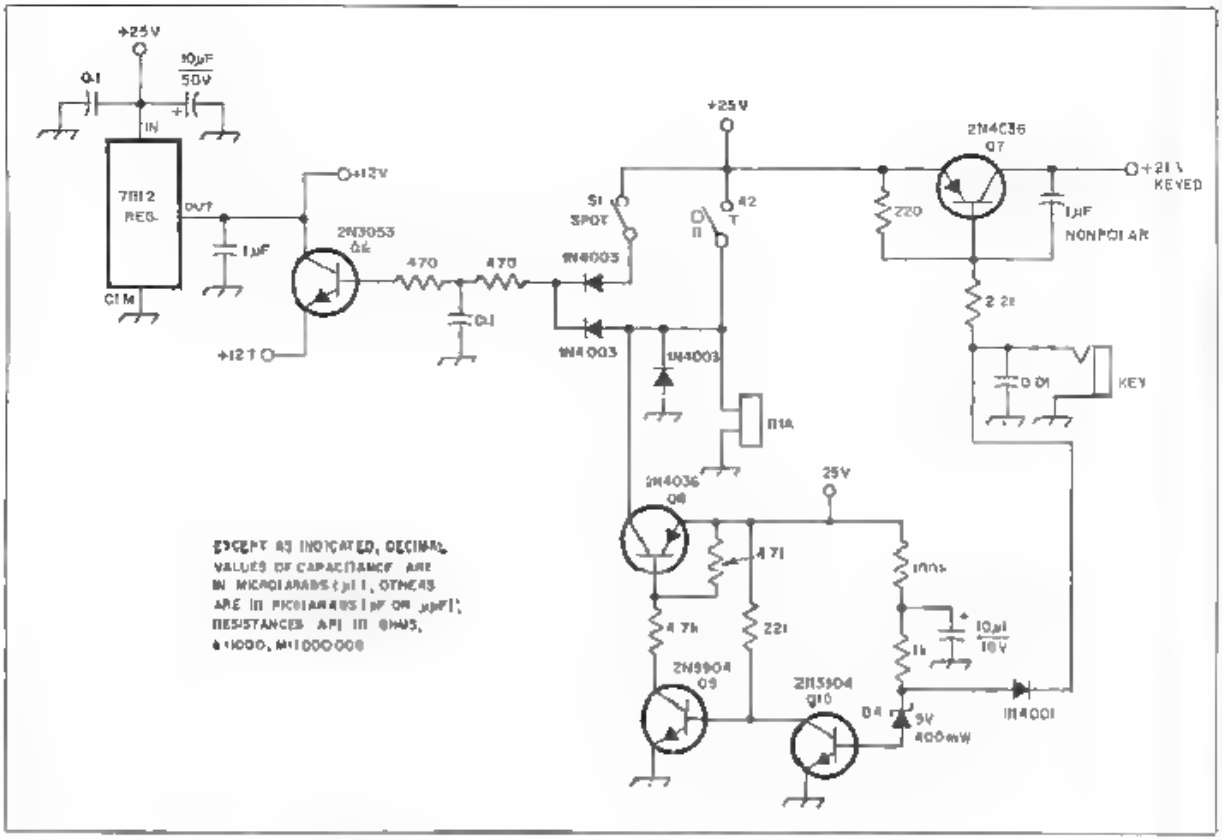

Fig. 2 — Control circuit for the VMOS FET transmitter.
K1 — 24 V dc dpdt relay. Magnecraft W88RPCX-3 or equiv.
D4 — 5-V 400 mW Zener diode, 1N757A or equiv.

driver stages were keyed, a requirement resulting from signal feedthrough in feedback amplifiers when they are "off," leading to an objectionable backwave.

A single band circuit was placed between Q3 and Q4. This improved the suppression of 14-MHz energy which was detected in the output of Q4 when using a 50-ohm termination and a 15-MHz-bandwidth oscilloscope. (Addition of the tuned circuit removed all traces of 20-meter energy from the oscilloscope presentation.) If the terminator is eliminated, the amplifier chain (Q3 and Q4) has a very wide bandwidth and is suitable for general-purpose application throughout the hf spectrum. A small heat sink is recommended for Q4.

The final amplifier is generally straightforward, with only a few subtleties. A low-pass filter is contained at the input. It serves the role of absorbing the input capacitance of the transistor, in this case about 50 pF, and hence aids broadband performance. The input of the amplifier is terminated in a 47-ohm resistor. While this decreases power gain, it does provide a low impedance at the gate, a definite aid to stability.

All presently available VMOS transistors are enhancement-mode devices. That is, with no positive voltage on the gate with respect to the source, there is no drain current. Only when a threshold gate voltage is reached does drain current begin to flow. This is typically +1.2 volts for the VN88AF. Current flow increases dramatically as the gate potential is increased further.

Many available VMOS FETs have an internal Zener diode connected between the gate and the source. The diode protects the FET from damage by static electricity. Only one Zener diode is used, in contrast to dual-gate MOSFETs, which employ back-to-back Zeners. The VN88AF includes a protection diode. If there were no internal protection diode, it would be possible to attach the gate directly to the 47-ohm resistor with no additional circuitry. But at rf, the negative-going portion of the rf voltage would quickly destroy the protection diode, taking the transistor with it. Hence, external protection circuitry is required to save the amplifier from the ill effects of the internal protection diode. The resistor network and 1N914 diode shown with Q5 serve this function, clamping the gate voltage and never allowing it to go below about 1.2 volts.

With no drive applied, Q5 sits on the verge of conduction. When drive is applied, the series 280-pF capacitance will charge, establishing a small positive dc voltage on the gate. As such, the amplifier operates Class A. The key-down drain efficiency is poor, only about 30 percent. For this reason, a husky heat sink is mandatory for Q5. (During testing, the VN88AF was destroyed from excessive dissipation because of an inadequate heat sink.) Overall efficiency is reasonable during typical cw operation, since forward bias disappears once drive is removed. Measurements have not been performed on this circuit when operated in a linear mode. However, the method might hold promise for ssb applications.

The output of the amplifier uses a double pi network. Following the work of Roy Lewallen,[7] such a network was used in anticipation of obtaining Zener-less devices that can be operated Class C or D. For the Class A operation employed in this design, a series-tuned output network would probably perform no problem.

The output power is +36 dBm, or about 4 watts. Slightly over 8 watts of output was obtained when a second VN88AF

was paralleled with Q5. No circuit changes were required other than retuning of the output network. Operation was attempted at a drain supply potential of 12 volts, but power output and gain suffered severely.

Some experiments that might be of interest were done on 80 meters. An amplifier much like that used at Q5 was built with a similar bias scheme. This amplifier used four paralleled VN88AFs bolted to a large heat sink. Power outputs up to 25 watts were easily obtained, but the efficiency was still poor. A similar 3.5-MHz amplifier was then built using a Siliconix VN84GA. This transistor is a real brute with no internal protection Zener diode. It had an output of over 25 watts with an efficiency of 73 percent. The amplifier was then moved to 14 MHz by retuning the 1000-pF gate capacitance. Similar results were obtained there. Unfortunately, this transistor is both expensive and difficult to obtain. Perhaps that situation will improve with time.'

The control circuitry for the transmitter is shown in Fig. 2. A 7812 three-terminal voltage regulator powers the low-level stages as well as a crystal-controlled receiving converter included within the same box. Transistor Q6 operates as a switch to apply voltage to the oscillator and buffer when either the spot or the transmit switch is activated. Q7 is a pnp switch controlled by the key to provide the voltage for Q3 and Q4. A 1-μF nonpolar capacitor from base to collector forces Q7 to act as an integrator during transitions. This shapes the keying nicely.

Transistors Q8, Q9 and Q10 form a semibreak-in circuit. When the key is pressed, the antenna relay is activated. It will remain on for a fraction of a second after the key is released. The transmit switch, S2, overrides the semibreak-in circuit for more casual contacts. If desired, Q8, Q9 and Q10 may be omitted. They were installed in this transmitter a few days before the annual November Sweepstakes contest. The antenna relay used was a surplus item from the junk box. There is nothing critical here.

The simplicity of the control circuitry presents one potential problem: The transmitter is on (and generating rf) at the instant the antenna relay changes to the transmit position. However, the low power and the inherent stability of the Class A final amplifier allow "hot switching" with no problem. Control systems for correcting this situation are described in chapter 7 of *Solid State Design for the Radio Amateur*.

Results

The performance of this transmitter has been as good as expected. Investigation with a Tektronix 7L13 spectrum analyzer *after* construction and alignment (using less exotic home-station test equipment) was encouraging. The 14-MHz component is 57 dB below the 28 MHz carrier. The second harmonic is 64 dB down while the fourth and sixth harmonics are just barely detectable. The backwave is over 75 dB down. The output amplifier has performed flawlessly with no sign of the usual instabilities found with bipolar power amplifiers. The VMOS FET power transistor is certainly here to stay![1]

On-the-air reports are equally encouraging. Keying and general "cleanness" are comparable to any of the better signals around. While using an inverted-V dipole only eight meters high this writer worked 41 states and a considerable amount of DX in the first two months of operation. The DX (in all continents) includes many slightly rare prefixes, ranging from LU and CX to HKØ and EA8. Let's just hope that the sunspots hold for several more years![2]

Notes

Rash, "MOSFET Power Amplifiers," *Ham Radio*, November 1978. Also see literature cited in that paper, especially those by Dorne.

Hayward and DeMaw, *Solid State Design for the Radio Amateur*, ARRL, 1977, p. 41.

Ibid., pp. 197–213.

Tweedie, "On Solid State PA Matching Networks," *QST*, October 1978, p. 38.

From February 1984 QST, p 35

A Beginner's Look at Basic Oscillators

A frequency generator is the heart of any signal source. Simple crystal or LC oscillators have many uses in amateur circuits. Let's learn how they work and where some common problem areas exist.

By Doug DeMaw,* W1FB

"Don't try to dazzle me with exotic circuits! I want to learn the theory of simple circuits first." Those statements are voiced frequently by radio amateurs. Are you one of those frustrated persons?

Perhaps the blind spot that exists with some writers (and I'm one myself) results from the belief that in order for a ham to have passed the license exam, he or she fully understood the answers to the theory questions. This is not a fact, because (unfortunately) many amateurs memorize the suggested answers to the FCC examinations. This makes it difficult to comprehend even the most basic of discussions about electronics.

Something else is awry for those who don't understand the fundamentals of our radio pastime. They can't experience the joys of building and using homemade gear! The purpose of this Beginner's Bench series is to encourage those of you who are less technically inclined to climb the ladder to a level that will enable you to enjoy the technical section of *QST* more fully, and to do some dabbling in your home workshops.

Perhaps the most common circuit in RF (radio frequency) projects is the oscillator. A single oscillator can serve by itself as a transmitter for CW. It may also be used as a frequency generator to be followed by one or more amplifier stages to provide a high-power transmitter. But, oscillators are used also in receivers, frequency standards, signal generators (test equipment) and many other pieces of apparatus for amateur use.

Perhaps you're saying to yourself, "Why hasn't he mentioned frequency synthesizers?" Well, that's not a topic that can be handled properly in a beginner's discussion. The synthesizer is a very exotic item that involves a host of subjects that are beyond the intent of this series. There's no doubt that synthesizers are becoming the way of life with most manufacturers of commercial amateur equipment. But, for the sake of experimenting with useful, simple circuits, we will focus on crystal and LC (coil and capacitor) oscillators. They are by no means obsolete!

What is an oscillator? In electronics, an oscillator is a device that generates an alternating current (ac). *Oscillation* is a variation in the magnitude of electrical current with time. Typically, the output of an oscillator alternates between positive and negative current values centered on zero current.

Everyone has alternating current available from an electrical wall outlet. Why is an oscillator needed to produce ac? The ac from the wall outlet is alternating at 60 Hz (cycles per second). In radio we need oscillators that will produce a wide variety of other frequencies from the audio range (20 to 20,000 Hz) throughout the radio-frequency range (as high as 300 GHz, or 300 billion cycles per second).

To make an oscillator, we must have two things. One is a frequency-determining element. This element is an energy-storage device with a special ability to build up energy in one direction, discharge it, build it up in the opposite direction, and discharge that. A pendulum is an example of a mechanical oscillator that does just that. Another example of a mechanical oscillator is the tuning fork used as a standard by musicians. Both of these mechanical devices store energy and oscillate at a certain frequency. In an electrical oscillator we generally use a quartz crystal or a tuned circuit consisting of a coil and a capacitor as the energy-storage and frequency-determining device.

The second ingredient of an oscillator is the ability to supply carefully timed pulses to keep it oscillating. Recall that a tuning fork oscillates for only a short while after it is banged against something. Similarly, a pendulum eventually winds down as the effects of gravity and friction win out.

Neither the mechanical nor electrical oscillators are perpetual-motion machines. The mechanical devices can be kept going by giving them a kick every now and then in exact timing needed to replace the power lost to gravity and friction. The same idea applies to electrical oscillators — there must be a pulse of electrical power supplied to the frequency-determining element exact-

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ly synchronized to the frequency of oscillation. The amount of power supplied must replace power lost to circuit resistance. This replacement power is called feedback. To obtain the extra power needed for feedback, it is necessary to sample some of the oscillating energy from the frequency-determining element, amplify it, and feed it back to the frequency-determining element so that it *aids* the power build-up. So an electrical oscillator needs an active device (such as a transistor or a vacuum tube) to serve as an amplifier to produce the correct feedback to keep the circuit oscillating.

Crystal Oscillators

A crystal oscillator circuit can be built with a quartz crystal and an amplifier to provide the needed feedback. When the amount of feedback is sufficient, the quartz element in our crystal holder will vibrate at a specified rate (depending on its thickness and the stray capacitances present in our circuit). The crystal is ground to the proper thickness at the time of manufacture, and the resultant frequency is marked on the crystal case. Therefore, if our crystal was marked "3 700 MHz," it would vibrate 3.7 million times a second to provide the desired oscillator frequency. The thinner the quartz crystal, the higher the operating frequency. This limits the practical upper frequency of a fundamental crystal, for if it were too thin the element would become impossible to fabricate or would shatter easily during oscillation. Generally, 20 MHz is the upper limit for quartz crystals that operate on their fundamental modes.

Although a crystal may be marked for a specific operating frequency, this does not mean it will produce that exact frequency when we plug it into an oscillator. The crystal must be ground or etched in accordance with the circuit capacitances that exist in our oscillator. This is specified by the manufacturer as the "load capacitance," the existing circuit capacitance that "loads" the crystal. Normally, the load capacitance of a standard oscillator circuit is somewhere between 10 and 40 pF, with 20 or 30 pF being the most typical value. Some circuits are very difficult to analyze with regard to the effective load capacitance. For this reason amateurs who need to have the crystal work at a precise frequency must tell the crystal supplier the model number of the equipment in which the crystal will be used. If the circuit is homemade, or if the model number is not known, the supplier should be provided with a copy of the oscillator circuit, with all parts values marked plainly on the diagram. We can take advantage of the effects of load capacitance by introducing changes in capacitance intentionally. This enables us to shift the operating frequency of a crystal. More on this subject later.

LC (coil/capacitor) frequency elements

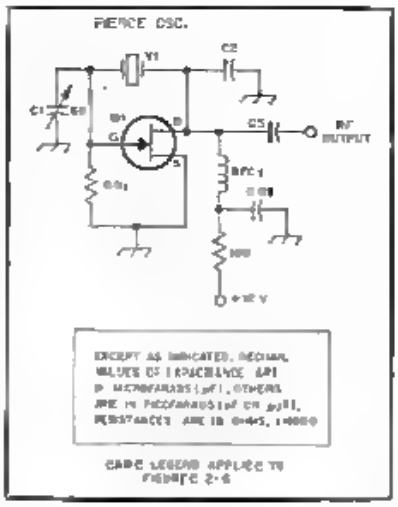

Fig. 1 — Simple example of a Pierce crystal oscillator.

in an oscillator must also be supplied feedback energy to cause oscillation. The coil and capacitor do not vibrate as is the case with a crystal element. Instead, the combination stores and discharges energy at a specific rate to establish the frequency of oscillation. The LC oscillator is seldom as frequency-stable as a crystal oscillator. Changes in temperature and mechanical vibrations (unwanted) tend to change the inductance and capacitance elements of the LC oscillator in a more dramatic manner than when a crystal is used. This causes an instant (mechanical) or gradual (electrical) change in the operating frequency. The gradual change is referred to as "drift."

The Pierce Oscillator

One of the simplest types we hams can use is the Pierce oscillator (named after a person, as are most oscillator circuits). Very few parts are required, as shown in Fig. 1. It makes no difference whether we use a vacuum tube (triode), bipolar transistor or an FET (field-effect transistor) in the circuit. The operating conditions remain the same except for the dc voltages applied to the circuit. The tube would require filament voltage and a higher dc voltage.

Y1, the quartz crystal, is located in the feedback path (between the drain and gate of Q1) to ensure oscillation. We must be careful to make certain we have neither too little nor too much feedback. Insufficient feedback will prevent oscillation, or sluggish starting of the oscillator when operating power is applied. Too much feedback can cause unwanted "frigglies" (oscillations at other than the crystal frequency) or, as some call the condition, "squegging."

To have control over the amount of feedback in Fig. 1 we have added C1 and C2. C1 is variable (a trimmer capacitor) to permit adjustment of the feedback energy. Once the correct value of capacitance is

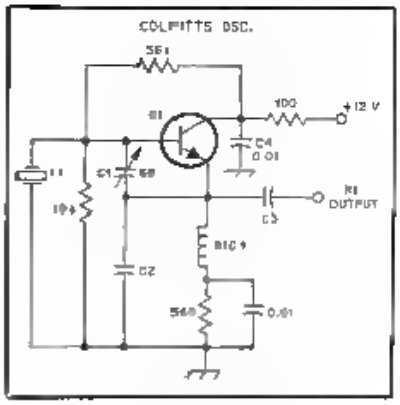

Fig. 2 — A basic Colpitts oscillator using a crystal.

found for our crystal, by virtue of C1, we may install a fixed-value capacitor. A 100-pF capacitor is suitable for C2 for ham-band use from 1.8 to 21 MHz. C1 can be a 60-pF trimmer. An MPF102 or 2N4416-family FET will be suitable at Q1. RFC1 is an RF choke that is resonant with the stray circuit capacitance (roughly 10 pF in most cases) well below the crystal frequency.

For example, using the ARRL Type A L/C/F slide-rule calculator, we would find that a 150-μH RF choke with 10 pF of stray capacitance would be resonant at the high end of the 75-meter band (4 MHz). If our oscillator were for use in that part of the spectrum we would want to avoid this condition. It would be better to use a 500-μH choke, which would provide resonance at approximately 2.2 MHz. We could remove all doubt by using a 1-mH (millihenry) RF choke, which is 1000 μH.

Output from the oscillator of Fig. 1 is taken from the drain of Q1. In order to help prevent the circuit that follows our oscillator from impairing oscillations (loading the oscillator too heavily), a small value of capacitance is used at C3. It should be the smallest value that is practical for delivering the required power to the next stage of the overall circuit. Usually, this will be between 50 and 100 pF in the 1.8 to 30 MHz range. Too much oscillator loading can prevent oscillation.

Colpitts Oscillator

A popular oscillator is shown in Fig. 2. This is the Colpitts circuit. Although a bipolar transistor is shown at Q1, a tube or FET could be used with equal success. In Fig. 1 we found the source of Q1 at ground potential, respective to dc and RF. In Fig. 2, the collector of Q1 is at RF ground by virtue of the collector bypass capacitor, C4. Hence, the feedback path for the Colpitts circuit we have illustrated is between the emitter and base. Other forms of the Colpitts oscillator are common, thus is but one variation.

Once again we have used two capacitors

(C1 and C2) for controlling the feedback. C1 and C2 are for that purpose. I find that in a practical circuit that uses a good, active crystal, the ratio of capacitance for C1 and C2 is on the order of 4:1. The larger value is used at C2. By placing a trimmer at C1 we can adjust the feedback for best performance of the crystal we use at Y1. A value of 100 pF seems to be fine for C2, with C1 being a 60-pF trimmer. The RF-choke rule for Fig. 1 does not apply here, entirely. The self-resonant frequency should be well below the crystal frequency. But, with a 100-pF capacitor in shunt with the choke (RFC1 and C2), the resonant frequency will be rather low compared to what it would be if only 10 pF of stray capacitance were present.

Output is taken via C3, which should again be a small value of capacitance to prevent the succeeding circuitry from loading the oscillator excessively. The 10- to 100-pF range is applicable to this circuit also. Q1 can be any small-signal bipolar transistor that has a fairly high cutoff frequency (f_T). I like to use a transistor that has an f_T of 5 to 10 times, or greater, the crystal frequency. Such devices as the 2N3904 and 2N2222A are fine for frequencies up to 20 MHz — the approximate limit for fundamental-cut crystals.

There are, indeed, many kinds of crystal oscillator circuits, but it would take many articles of this length to show them and describe their basic performance characteristics. The Pierce and Colpitts form the basis for most amateur oscillator circuits.

Overtone Oscillators

How might we obtain crystal-oscillator performance above the frequencies for which fundamental crystals are limited? Well, we adopt what is called the "overtone oscillator." As is true of fundamental types of oscillators, there are countless overtone-oscillator circuits. We will deal with but two of them, mainly to illustrate the principle of operation. A simple triode overtone oscillator is shown in Fig. 3A. Y1 is manufactured as a crystal that operates at an odd multiple of its fundamental frequency. This means that we may use a third- or fifth-overtone crystal in our circuit to obtain output at some frequency above, say, 20 MHz. Let's imagine that we wanted a crystal oscillator for use at 28 MHz. We should order a third-overtone crystal for the exact 10-meter frequency of interest. The manufacturer would again need to know the load capacitance presented by our circuit in order to grind for each quartz correctly. The crystal is ground for roughly one third the operating frequency. That is, a 28-MHz crystal would be ground for approximately 9 333 MHz. An overtone crystal does not oscillate at exactly three times the frequency of the quartz element, however, so the manufacturer must know the exact overtone frequency we desire. Likewise with fifth-overtone crystals, and so on.

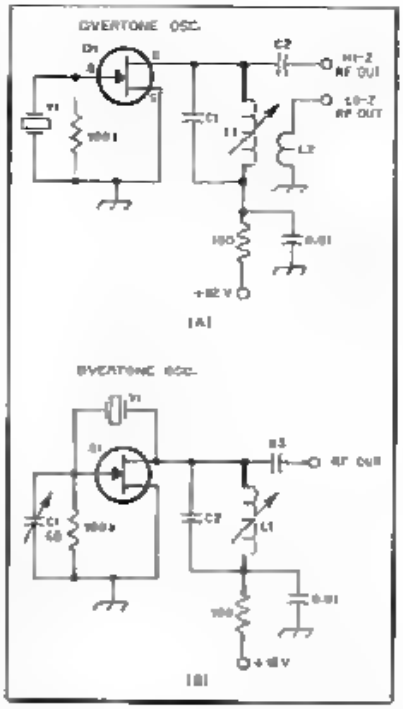

Fig. 3 — Two common types of crystal overtone oscillator.

The circuit at A of Fig. 3 is rather simple. Sufficient internal coupling exists within Q1 to provide the feedback we need for oscillation. This would not necessarily be true of oscillators operating at the fundamental mode of the crystal. C1 and L1 are tuned to the desired overtone frequency, thereby providing feedback at the required frequency. If all is as it should be, Y1 will oscillate and provide RF output from Q1 at *only* the overtone frequency. Too much feedback will permit the crystal to oscillate at its fundamental frequency. This will cause the oscillator output to contain two frequencies — the fundamental *plus* the overtone. Output can be taken at high impedance by means of C2, or a link can be wound on L1 to provide low-impedance output via L2. The choice will depend on what we couple our oscillator to.

Another kind of overtone oscillator is illustrated at Fig. 3B. At first glance we might conclude that it is a Pierce oscillator. But, it is an overtone type of oscillator, with the crystal inserted between the drain and gate of the FET. C1, used to control the feedback, will have a slight effect on the operating frequency as it is adjusted. C2 and L1 again form a resonant circuit at the overtone frequency.

LC Oscillators

Most LC oscillators are used as VFOs (variable-frequency oscillators). But, we may elect to use them on occasion as single-frequency devices, just as we would with a crystal oscillator. How useful an LC oscillator may be will depend entirely on

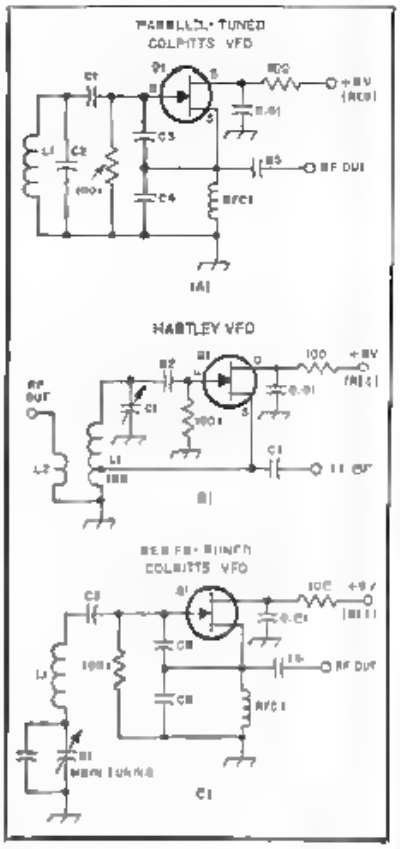

Fig. 4 — Examples of coil-capacitor oscillators (see text).

how frequency-stable we can make it. Although crystal oscillators are more expensive, they do offer the best stability of the two types.

Acceptable frequency stability is obtained through careful selection of the circuit components, the amount of feedback used, regulation of the operating voltages and providing as nearly a constant temperature environment as possible. Special temperature-compensating capacitors are often used to minimize frequency changes. The coil and capacitor must be mechanically and electrically well built to enhance stability. Similarly, nothing in the immediate vicinity of the LC oscillator must be allowed to move position, for this can change the operating frequency. Changes in oscillator loading, caused by operating-condition variations in succeeding circuit stages, will also shift the frequency. LC types of oscillators are more prone to this malady than are crystal oscillators.

Three types of LC oscillators are shown in Fig. 4. The first example (A) is probably the most common of the three in ham equipment. Since C2 and L1 are in parallel, this is called a parallel-tuned oscillator. C3 and C4 provide the path for our feedback energy. In LC oscillators the value of C3 and C4 are approximately the same. A

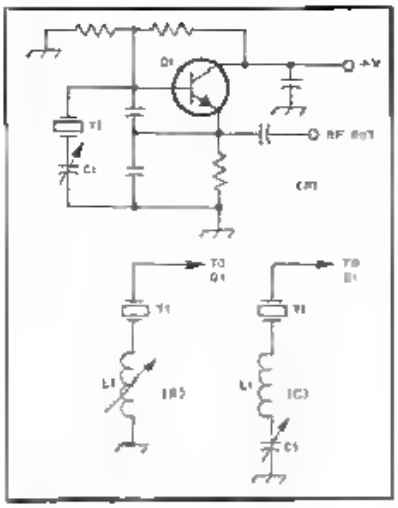

Fig 5 — Details of how C, L or both elements can be added to an oscillator circuit to modify the crystal frequency.

3.5-MHz VFO, for example, might use 1000 pF for each of the feedback capacitors. C5 is once more a small-value coupling capacitor to minimize output loading.

The circuit of Fig. 4B relies upon a tap near the grounded end of L1 to provide feedback for oscillation. The tap is usually located approximately ¼ of the way up the coil from ground. Output can be taken via C3 or by means of a link winding (L2) at the ground end of L1.

A series-tuned Colpitts VFO is illustrated at Fig. 4C. The same applies because C1 and L1 are in series for this circuit. The advantage of this configuration over that of Fig. 4A is of particular interest as we raise the operating frequency. Sometimes we end up with impractical (small) values of inductance for L1 in a parallel-tuned circuit, especially at high operating frequencies. The series-tuned arrangement requires a much larger amount of inductance and a smaller value of capacitance at C1 for a given tuning range. The series-tuned format can often lead to improved frequency stability for reasons that we cannot address here.

Shifting the Crystal Frequency

We discussed earlier the possibility of making small changes in the operating frequency of crystal oscillators. This can be done by adopting the methods depicted in Fig. 5. The smallest frequency change will be experienced when using only a variable capacitor as shown at A in Fig. 5. The higher the crystal frequency the greater the frequency shift as C1 is adjusted. The trimmer can also be placed in parallel with C1. One method raises the frequency while the other lowers it. An inductance can be added in series with a crystal, as shown in Fig. 5B, to shift the frequency of oscillation. Too much inductance at L1 will convert the crystal oscillator to an LC type of circuit, and the benefits of crystal control will be lost. I prefer to use an inductive reactance of approximately 850, maximum. Hence, for a 7.0-MHz crystal the maximum inductance at L1 would be 19.3 µH, derived from

$$L(\mu H) = \frac{X_L}{2\pi f} = \frac{850}{6.28 \times 7} = 19.3 \quad \text{(Eq. 1)}$$

where X_L is in ohms and f is in MHz.

The greatest amount of frequency pulling or swing will be obtained if we employ the method at C of Fig. 5. Here we have a coil and capacitor in a series arrangement at the bottom end of Y1. A 100-pF variable capacitor can be used along with a coil whose value is derived from Eq. 1. Frequency shifts as great as 10 kHz can be had at 10.1 MHz, with 5 kHz being typical at 7 MHz, and 3 kHz being about the limit at 3.5 MHz. Anything greater than that suggests that L1 has too much inductance for full crystal control. A circuit like the one in Fig. 5C is usually referred to as a VXO (variable crystal oscillator). In some circuits we will find that C1 has been replaced by a varicap diode, or voltage-variable capacitance diode. The frequency change will not be as great as with an air variable capacitor, since the minimum capacitance of a varactor diode will be much higher than that of a mechanical capacitor.

Buffering and Isolation

Throughout our discussion we have mentioned loading at the output of oscillators, plus the frequency shifting caused by load variations. We considered also the effects on oscillation that too much loading might cause. These problems can be reduced or eliminated by adding buffer stages after the oscillator, as shown in Fig. 6. In effect, these additional stages help to isolate the oscillator from the circuits that succeed the frequency-generating chain. Some buffer stages can also provide signal amplification, whereas others might reduce the effective output level of our oscillator. FETs work well as buffer stages, owing to their very high input impedance (usually a megohm or greater). The gate resistor in Fig. 6 determines the input impedance of Q2, since it is lower in ohmic value than the natural gate impedance of Q2. Since we show Q2 as a source-follower stage, the output of the FET will be slightly less than the output of Q1 — approximately 10% lower.

Most VFO circuits have at least two buffer stages, and sometimes three. One or more of the buffers can be designed as amplifiers if we wish. This enables us to extract greater output power than would be possible if we took the output directly from the oscillator. C1 and C2 of Fig. 6 are small in capacitance value. This helps limit loading effects after the oscillator. If you have built a VFO-controlled CW transmitter that is chirpy (frequency shifting when the key is closed), chances are that you did not include sufficient buffering to isolate the VFO.

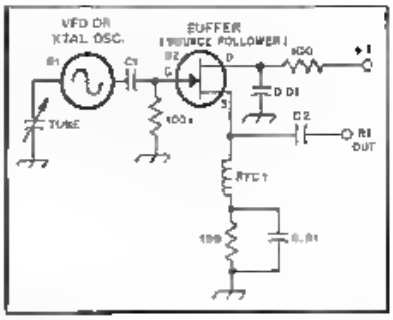

Fig 6 — Illustration of how a buffer stage can be used after an oscillator to minimize frequency changes caused by load variations.

A Practical Universal VXO

I lean rather strongly toward the use of VXOs (Fig. 5C) above 7 MHz, especially for portable transmitters and receivers that are apt to be used in an environment of frequent temperature changes. They are stable and reliable. The VXO is nice for use as a VFO when operating VHF equipment. We will not obtain as great a potential frequency swing with our VXOs as can be had with an LC type of VFO, but more than one crystal can be switched into the VXO for wide frequency coverage in some amateur bands.

The circuit in Fig. 7 shows the diagram of a VXO I developed for my use on a number of frequencies. C1, C2, C3, FL1 and Y1 can be changed to appropriate values for the frequency of interest. This circuit is set up for use as a 2-meter VFO, and when its output is multiplied eight times to 144 MHz, I can obtain coverage from 144 to approximately 144.250 MHz — about right for the CW and SSB part of the band.

Those of you who like to experiment may want to build this circuit. It can have many uses, depending on the frequency to which it is tailored. For example, we might use a VXO for the local oscillator in a homemade receiver. It could be the heart of a little signal generator for workshop use. We might multiply the output more than eight times for the purpose of using the VXO as a frequency source at 220 or 432 MHz, or as a signal generator for VHF and UHF testing. By lowering the VXO frequency to 20, 30 or 40 meters, it can serve nicely as the frequency-controlling element for a home-built CW transmitter.

Best operation (maximum frequency swing) will be had if we use AT-cut crystals. Preferably, they will be the type that are suspended by tiny wires inside the crystal holder (HC-6/U), and they will be cut for fundamental-mode use. I use International Crystal Mfg. Co. general-purpose types of crystals with a 30-pF load capacitance.

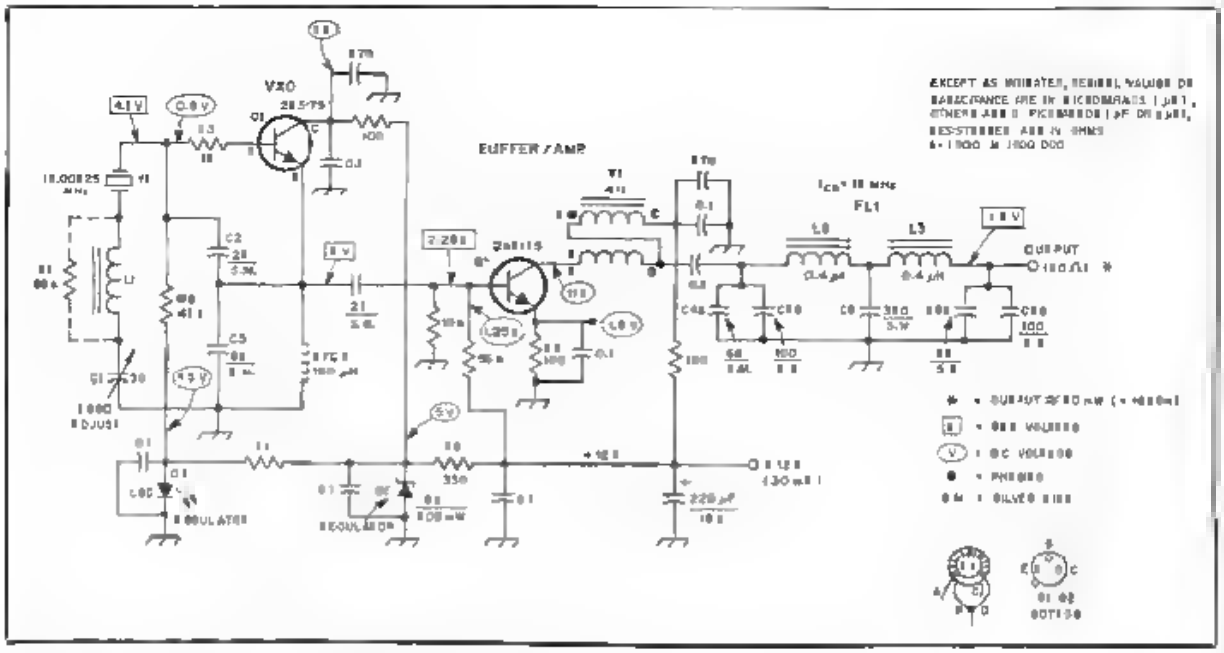

Fig 7 — Schematic diagram of a universal VXO. Fixed-value capacitors except those in Table 1 are disc ceramic. The polarized capacitor is tantalum or electrolytic. Resistors other than R3 are ¼- or ½-W carbon composition. R3 is a ½-W trim

C1 — Panel mounted miniature air variable 50 or 100 pF (mail critical).
C2-C6 incl. — See Table 1.
L1, L2 and L3 — See Table 1.
D1 — 1.5-V regulator. Use an LED or two 1N914 diodes in series.
Q2 — 8 V, 400 mW Zener diode.
R2 — Value may be adjusted for purest output waveform from Q1.
T1 — Broadband toroidal 4:1 transformer (see Table 1).
Y1 — AT-cut fundamental crystal in HC-6/U holder. 20- or 30-pF load capacitance suitable (International Crystal Mfg. Co. type GP).

Table 1

Typical L and C Values for Various Operating Frequencies

Crystal (Y1) Range (MHz)	C2 (pF)	C3 (pF)	C4 C5 (pF)	C5 (pF)	L1	L2, L3	T1
8 000 to 9 000	68	100	350 (for 40-meter use)	820	37 µH max. 55 ts. of no. 26 wire on a T68-2 toroid core	1 µH. 18 ts. no. 28 wire on a T37-6 toroid core (40-meter use)	16 bifilar ts. of no. 28 on an FT-37-43 toroid core
8 000 to 15 000	36	68	287 (for 30-meter use)	560	12 µH max. 48 ts. of no. 28 wire on a T50-2 toroid core	0.72 µH. 15 ts. no. 28 wire on a T37-6 toroid core (30-meter use)	16 bifilar ts. of no. 28 on an FT-37-61 toroid core
8 000 to 16 000	39	68	212 (for 20-meter use)	424	12 µH max. Same as above.	0.53 µH. 13 ts. no. 24 wire on a T37-6 toroid core. (20-meter use)	Same as above on FT-37-61 toroid core
15 000 to 20 000	27	56	166	330	7 µH max. 42 ts. of no. 26 wire on a T30-6 toroid core	0.42 µH. 12 ts. no. 26 wire on a T37-6 toroid core (for 15-MHz use)	Same as above on FT-37-61 toroid core.

The capacitors are silver mica or polystyrene types. All toroid coils should be dipped with two coats of coil cement or General Cement G-Dene® after they are wound. Toroid cores are available by mail from Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607. Also stock QST ads for Palomar Engineers and RadioKit. Values for C4, C5, C6, L2 and L3, when FL1 of Fig. 7 is to be used for frequencies other than those listed, can be obtained from the filter tables in *The Radio Amateur's Handbook*, transmitting chapter. Nonstandard capacitor values can be closely approximated by using series or parallel combinations of standard values. The nearest standard values have been listed where possible.

They are the least expensive and seem to be very "rubbery" in VXO circuits. Irrespective of the brand of crystal used, no two identical crystals will yield the same amount of frequency swing in a VXO. I have never understood exactly why this happens, but I have observed it countless times.

If a coil with very high Q (quality factor) is used at L1 of Fig. 7, it may be necessary to swamp the coil with a resistor (R1) to lower the Q. Values from 10 kΩ to 27 kΩ seem to do the job. The need for R1 will be evident if while we are tuning C1 through its range the VXO will break into oscillation at some other frequency (mode changing) and be erratic in operation.

As an aid to the overall frequency stability of Q1, I decided to regulate not only the collector voltage, but put a separate regulator on the feed for the base bias. D1, a standard LED (light-emitting diode), makes a fine 1.5-V regulator when used as shown. Alternatively, we can use two 1N914 diodes in series at D1.

Q2 functions as a buffer/amplifier. It is a broadband, linear Class A stage. Output from Q2 is filtered by means of FL1, a half-wave harmonic filter. It is designed for a 50-ohm input/output characteristic.

Typical output power is 40 mW, which equates to +16 dBm. This is ample to excite most low-level amplifiers or DBM (doubly balanced mixers) of the diode-quad type. If the DBM calls for −7 dBm of injection, a resistive 50-ohm attenuator can be inserted in the line between T1 and the mixer. The *ARRL Electronics Data Book* (out of print) contains tables of values for resistive attenuator pads. Circuit boards and complete parts kits for this workshop project are available.[1]

I have included key dc and RMS (root-mean-square) voltages on the schematic diagram. These can be used for trouble-shooting the circuit if problems arise. An RF probe and VTVM can be used to check the RMS voltage values, or you can use a scope if it has ample bandwidth to provide accurate P-P (peak-to-peak) voltage readings. Multiply your P-P voltages by 0.3535 to obtain the equivalent RMS voltage. All *dc readings are referenced to circuit* ground.

Although 2N5179 CATV transistors are specified in Fig 7, other NPN devices of similar characteristics will work satisfactorily. I have used 2N3572s with good results. The common 2N2222A should offer acceptable performance as well. The output power of this circuit can be lowered by increasing the resistance of R3. This can eliminate the need to add outboard attenuator pads for power reduction.

If possible, use a double-bearing variable capacitor at C1 (a bearing at each end of the rotor). Addition of a panel-mounted vernier drive will facilitate dial calibration and provide a better (slower) tuning rate. A frequency counter can be used to develop a dial-calibration chart.

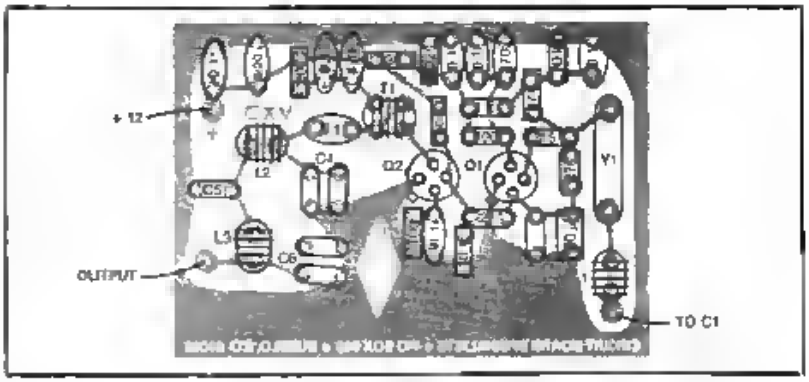

Fig 6 — Parts-placement guide for the universal VXO shown from the component side of the PC board.

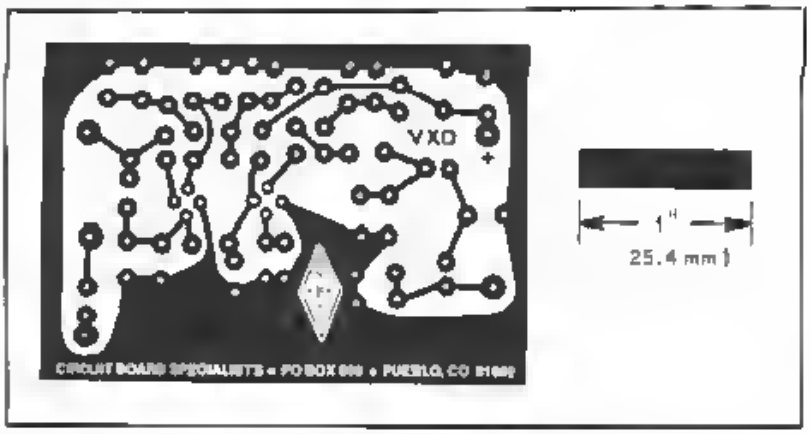

Fig 8 — Circuit-board etching pattern for the universal VXO. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil.

Frequency drift from a cold start to an hour later (at 70° F) was 30 Hz. At 2 meters this would multiply to 240 Hz — an acceptable value. The VXO or any LC VFO should be built in a separate shielded box for best results. This will help to maintain a more constant temperature and will prevent unwanted RF energy from entering the circuit and causing frequency changes that aren't wanted. Table 1 lists some typical values for other operating frequencies. Fig. 8 shows the parts placement for the circuit board. A scale pattern for the PC board can be found in the Hints and Kinks section of this issue of *QST*.

Closing Thoughts

We have barely scratched the veneer in this discussion of oscillators. But, I hope you have acquired a better understanding of how they work and what can be done to improve their performance. I suggest you take soldering iron in hand and tack together some of the one-stage oscillators that are presented in this article. Experiment with them to study the cause and effect of value changes, and so on. There is no substitute for "learning by doing." There is no reason you can't tackle the VXO project of Fig 7. It can be useful in many applications in your ham shack. Good luck!

[1]Circuit Board Specialists, P O Box 969, Pueblo, CO 81002. Catalog of kits available on request.

For updated supplier addresses, see ARRL Parts Supplier List in Chapter 2.

From July 1985 QST, p 22:

The Fine Art of Improvisation

Improvising in the ham workshop may lead to new ways for solving electrical and mechanical problems. The net result is often a savings in time and money!

By Doug DeMaw,* W1FB

"I gave up on building ham gear because parts are hard to find and they cost too much." Ever hear that comment? Perhaps you've said it to yourself in silent despair. Actually, parts are not hard to find, and most of them need not be purchased at top price. But, there are *some* items that are very expensive and hard to locate when we attempt to buy them new. It is conceivable that we might have to spend $15 for a tuning capacitor and a vernier drive, when the circuit with which it will be used contains only $3 worth of small parts. Prices of items such as tuning capacitors, drive mechanisms, cabinets, slug-tuned coils and meters (purchased new at nonsurplus prices) can discourage even those builders who have a large Amateur Radio budget. The cost, plus the present-day rigorles of being socked with back orders and "out of stock" notifications from mail-order dealers, does tend to make us think parts are hard to obtain.

What alternatives do we have? The ingenuity of a true experimenter must be summoned from within if speedy solutions to these common problems are to be found. In decades past it was a regularly practiced art among hams to solve design and procurement problems by using materials on hand. Most hams were inveterate experimenters when I became involved in Amateur Radio. It was considered a challenge to come up with new electrical and mechanical ideas, then thrill them with other amateurs. In those days, it was often a stimulating learning experience to get on the air and talk about circuits and projects. Each of us has the potential to build radio equipment, to find shortcuts to design objectives and to enjoy using something we built ourselves. Let's consider some practical ways to use parts in applications for which they were not designed. Perhaps some of these concepts will solve a design problem for you.

Experimental Tuning Methods

Transmitters and receivers require some type of signal source, and generally this local oscillator (LO) is tunable. The conventional techniques for changing a VFO frequency are by means of a fixed-value inductor and variable capacitor or a fixed-value capacitor and a variable inductor, or by employing a VVC (voltage-variable-capacitor) diode. A quality double bearing tuning capacitor that rotates smoothly is not only hard to find these days, it can be bulky and very expensive. Much of our miniature homemade equipment would be made practical if a tuning capacitor could be avoided for changing the oscillator frequency.

How might we contrive a simpler, less expensive method for tuning a VFO? I developed an interesting circuit for use in a very compact receiver (Fig. 1) that qualifies as a simple, inexpensive tuning technique. I had some reservations about

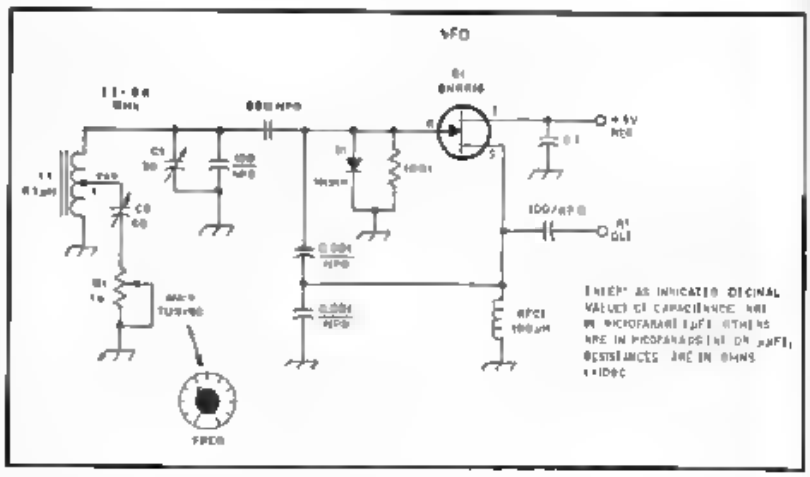

Fig 1—Typical circuit for a VFO that uses a 2N4416 or MPF102 FET. Tuning is by means of RI in series with C2. C2 sets the frequency spread provided by RI. This arrangement is useful when an air-variable capacitor and vernier drive are not desired. It can lead to a very compact VFO assembly. The tap position on the coil (L3) and the maximum capacitance provided by C2 determine the maximum tuning range available.

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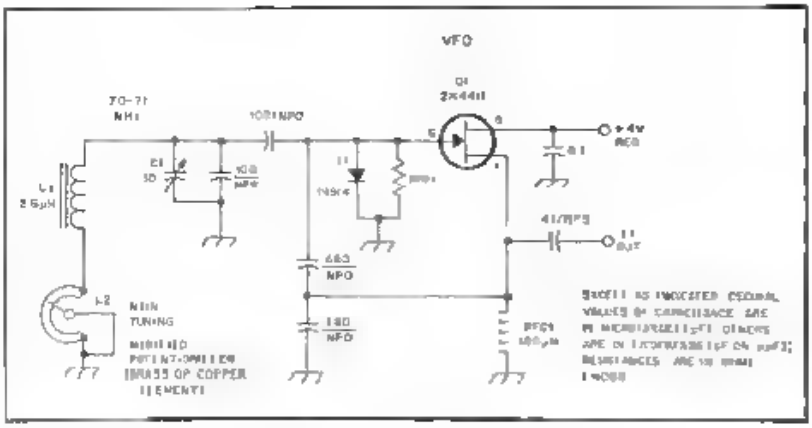

Fig 2—A variation of the circuit of Fig. 1. In this example, the tuning mechanism is a potentiometer that has been modified to become a small variable inductor (see text).

how it might work, but after breadboarding a test circuit, I was pleasantly surprised with the results. For lack of a better name, I call it "reactance tuning." Fig. 1 shows the details of the test circuit in which I tried the idea. R1, which is a high-quality Allen Bradley (A/B) potentiometer, is located close to C2 and L1 in order to keep the leads from R1 as short and direct as possible.

Why does this system work? Well, as R1 is adjusted, the presence of the capacitance of C2 (a trimmer) is more prominent in the tuned circuit. The series combination of C2 and R1 form a capacitive reactance and resistance that cause a frequency shift as R1 is adjusted. The smaller the value of resistance at R1, the lower the operating frequency, because the capacitance of C2 will be more effective.

What are the bad features? No innovation is necessarily perfect, and this applies to the technique illustrated in Fig. 1. The tuning is nonlinear. That is, the frequency is spread out at the maximum-resistance end of the R1 range, and it is somewhat compressed at the minimum-resistance end. Also, if a poor-quality control is used at R1, you may hear a slight scratching noise as the control is adjusted, while listening to the output of a receiver in which this VFO is used. It should not cause a problem if we use the VFO in a transmitter.

The amount of frequency shift available depends on two things: the position of coil tap X on L1 of Fig. 1, and the setting of the trimmer capacitor, C2. The farther the L1 tap is above ground, the greater the frequency change as R1 is adjusted. Similarly, the greater the capacitance of C2, the larger the frequency change. I had no trouble covering all of the 40-meter band when the coil tap was close to the high end of L1. In a practical application, it is best to limit the frequency change to 25 or 50 kHz. This provides better bandspread when R1 is adjusted. A vernier-drive mechanism can be coupled to R1 if frequency excursions greater than, say, 50 kHz are desired.

I did not observe any degradation in VFO frequency stability when comparing this tuning method with that of variable-capacitor tuning while using the same oscillator module. There is, however, a point in the tuning range of R1 where the loaded Q of the VFO tuned circuit will take a dip. When this happens, the VFO output will drop slightly and the output waveform linearity will change. In most practical applications, you will not be able to detect this effect.

As an alternative to the use of a vernier drive attached to R1, we might consider using a bargain-priced 10-turn, carbon-composition control with a suitable 10-turn counter dial. Wire-wound controls must be avoided because they are inductive.

Another Tuning Trick

I have another idea that I had in mind for a number of years. The circuit for this one is given in Fig. 2. L2, a modified carbon control, is fashioned by removing the metal cover from a standard-size potentiometer, then removing (carefully) the semicircular carbon element from inside the control. I was able to snap this element loose by prying it up near the tabs of the control. The thin phenolic base material broke easily. I used this element as a pattern and cut out a new element from flashing copper. Brass would work, also. Silver plating would help to ensure minimum corrosion, but it is not necessary to add silver plating if the control will be used regularly.

The new element is glued in place, and the ends of the insert piece are soldered to the two outer lugs of the old control. Be careful to avoid getting epoxy glue on the upper surface of the metal element, or erratic operation will result.

Refer again to Fig. 2. L2 is a small variable inductor we made from the potentiometer. It comprises a part of the overall circuit inductance by virtue of its being in series with L1. As L2 is adjusted, the VFO frequency will change. The higher the operating frequency of the VFO, the greater the frequency change caused by L2. Also, the higher the C-to-L ratio of the VFO, the more effect you will observe when L2 is adjusted. The frequency shift obtained with this method is substantially less than with the circuit of Fig. 1, at least with the circuit values given. A 10-kHz shift was observed.

Incremental band-segment selection can be had with either circuit (Figs. 1 and 2) by adopting the method shown in Fig. 3. S1 is used to add capacitors to the VFO tuned circuit, and R1 or L2 can be used in the manner described previously. Perhaps a miniature DIP switch can be added to operate S1 when compact equipment is being built. The values of capacitors C3, C4 and C5 will determine the coarse tuning range. Trimmers may be substituted for these fixed-value capacitors, which will enable you to have the tuning ranges overlap.

Simple, Homemade Tuning Capacitor

Large frequency changes are possible if we use a low-capacitance variable capacitor that is connected to the high end of a VFO tuned circuit (junction of C1 and L1 of

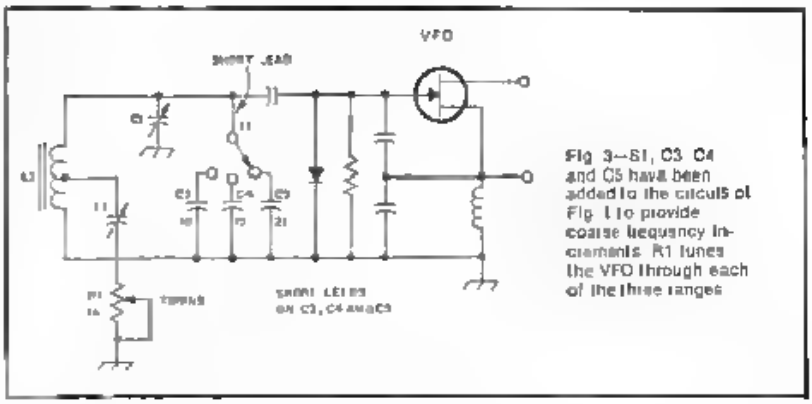

Fig 3—S1, C3, C4 and C5 have been added to the circuit of Fig. 1 to provide coarse frequency increments. R1 tunes the VFO through each of the three ranges.

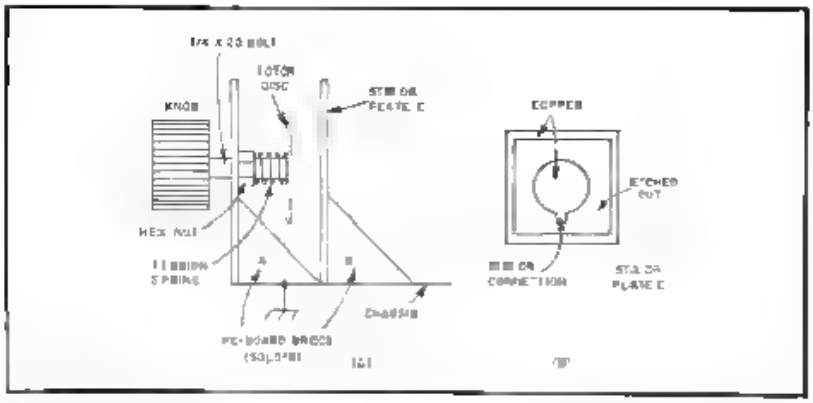

Fig 4—Mechanical details for a homemade disc tuning capacitor. A tension spring ensures mechanical stability of the rotor portion of the variable capacitor. Side brackets also help to keep the unit mechanically rigid. This detail at B shows how the stator disc is etched on PC board material.

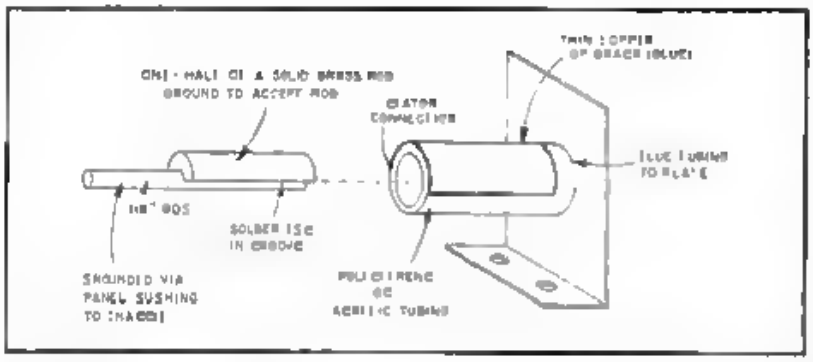

Fig 5—A cylindrical format provides still another tuning device that can be made at home. The rotor unit is semicircular brass or copper to which a ¼-inch-diameter tuning rod has been soldered. The stator section is a piece of plastic tubing to which thin copper or brass sheeting has been glued (see text).

Fig. 1). A simple mechanism is illustrated in Fig. 4. It is one that I developed during my search for simple VFO tuning methods. The drawing at A of Fig. 4 shows a side view of the assembly I constructed. A piece of ¼-in × 20 iron bolt is used as the tuning shaft. The front plate of the inner is a piece of copper-clad PC board. The hex nut is soldered to the inner surface of this end plate, as shown. A disc of copper or brass serves as the capacitor rotor. It is soldered to the end of the bolt that is opposite the knob. I used a 1-inch-diameter disc, and made certain it was at an exact right angle to the bolt when I soldered the two pieces together. A spring is used between the disc plate and the front-plate bearing nut to prevent wobbling and undue backlash. PC-board braces are soldered (four each) to the front bracket and stator plate bracket to ensure physical stability.

Drawing B of Fig. 4 shows how I made the stator plate. It is a piece of PC board with an outer border and disc that were provided by etching with ferric-chloride solution. Glass-epoxy circuit board is recommended in the interest of high dielectric quality and physical strength. A piece of thin Teflon® sheet is glued to the surface of the stator disc to prevent short circuiting of the stator and rotor discs. Polyethylene sheeting is suitable if you have no Teflon on hand. The capacitance range I obtained with this unit was 0 to 18 pF with the 1-inch diameter disc plates. The closer the plates are to one another, the greater the capacitance and the faster the tuning rate. The rotor disc is grounded by means of the bolt-to-nut connection and by virtue of the front-end plate being grounded. Those skilled in machine work should be able to improve on this design. The disc-tuning method is by no means a new concept. VHF cavities and amplifiers were tuned by this technique for many years. But, I don't recall seeing it applied to HF circuits to this manner.

Another Capacitor Idea

A cylindrical tuning capacitor can be fashioned as shown in Fig. 5. The rotor is slipped inside the stator tubing. When the metal half-rod of the rotor is immediately adjacent to the metal half-round outer conductor of the stator tube, maximum capacitance will exist. The rotor shaft is rotated by means of a knob or vernier drive to operate this capacitor. The larger the two half-round conductors (circumference and length), the greater the maximum capacitance of the unit. The mechanical aspects of this device can be improved markedly by those of you who are adept at building mechanical gadgets. Certainly, a fine assembly could be turned out by a craftsman. The point being made here is that this is just another method for constructing a homemade variable capacitor. There are many other unique ways to construct home-built tuning capacitors, but we shall not go into a lengthy discussion about them.

Generating Innovative Ideas

I have been asked, "How do you come up with so many unusual gadgets?" I think the best reply I can offer is to say that examination of a conventional component should suggest numerous ways to simplify it at a savings in cost. Some inventors do not generate new ideas. Rather, they pick up some ordinary object, such as a paper clip, then ask themselves, "What can I do to improve this thing?" We might also ask ourselves, "What don't I like about this paper clip?" The next step is to devise a new paper clip that no longer has the design fault. Bingo! A new patent can result! This general philosophy can be applied to making our own radio components from readily available materials. You can try your ideas, and you need not be ashamed if they don't work the first time or even at all.

In Conclusion

When you are working with the circuits of Figs. 1 and 2, it is important that the Q of L1 be as high as you can make it. If the Q is quite low, the addition of R1 or L2 could cause the VFO to cease oscillating at some point in the tuning range. Therefore, I suggest that you use a T68-6 toroid core for frequencies above 4 MHz. The wire size should be as large as can be wound easily on the toroid core. This will reduce the coil resistance and enhance the Q. The same rule applies if you use a slug-tuned inductor. The core should be for use in the upper part of the HF spectrum. High-quality capacitors should be used also. The NP0 units specified are entirely suitable, and will ensure minimum VFO drift. Silver-mica capacitors can be used, but will cause considerably more drift than will the NP0 ceramic units.

Should you develop some noteworthy circuit innovations, please consider sharing them with others through the pages of *QST*. Detailed descriptions can be submitted as articles. Short explanatory narratives may be just right for the Hints and Kinks column.

From September 1987 QST, p 25:

Tuning-Diode Applications and a VVC-Tuned 40-m VFO

Inexpensive voltage-variable capacitance diodes are compact and easy to use in your Amateur Radio circuits. They can replace expensive, hard-to-get air-variable tuning capacitors.

By Doug DeMaw W1FB
ARRL Contributing Editor
PO Box 250, Luther, MI 49656

Have you looked lately for small air-variable capacitors? Does the high cost and scarcity of tuning capacitors bother you? If you answer "yes" to these questions, I can sympathize with you!

Gone are the days when small air-variables lined the shelves of electronics surplus stores. Gone, too, are the attractive price tags of $1 or less. New capacitors are presently in the $10-$15 class, if you can find them. The once-popular Hammarlund and E. F. Johnson capacitor lines are produced by another firm, and single-lot purchases are a thing of the past. The surplus market has literally dried up for small air-variable capacitors with tuning shafts. There is, however, a bright spot in this seemingly grim situation.

We can look toward voltage-variable capacitance (VVC) diodes as a solution to the mechanical-capacitor shortages, at least for use with low-power oscillator and low-level tuned RF circuits. Tuning diodes are not only inexpensive, they are small. There is a greater opportunity for circuit miniaturization using VVC diodes. The major performance trade-off relates to use of diodes in VFO circuits: The frequency stability may be worse than with air variables, and the minimum capacitance of VVC diodes is substantially greater than is typical of an air-variable capacitor. For most amateur applications, however, these shortcomings are not serious.

VVC Diode Characterization

You have probably heard people refer to tuning diodes as Epicap® or Varicap® diodes. These are trade names that the manufacturers have given to these diodes. A varactor (variable reactor) diode is similar in effect to a tuning diode, but it is earmarked for use as a frequency multiplier (harmonic generator). Ordinary tuning diodes work quite well as varactors, as do many small-signal, high-speed switching diodes, such as the popular 1N914. The base-collector junctions of many transistors may also be used as tuning diodes or varactors.

In simple terms, the junction capacitance of a VVC diode changes when a reverse voltage is applied to the device (positive voltage applied to the diode cathode), and the capacitance varies with the voltage. The diode is placed in parallel with the components of a tuned circuit, and tuning is accomplished by varying the voltage, and thus the capacitance, by means of a potentiometer.

Fig 1 shows the equivalent electrical circuit of a VVC diode. Note that there are components of capacitance, resistance and inductance present. C_c is the stray case capacitance. C_j is the junction capacitance (voltage variable). L_s is the diode series inductance, and R_j is the junction resistance (also voltage variable, but negligible above 100 kHz). Finally, R_s is the series resistance of the diode and its leads. Our practical concern is mainly for the C_j component, at least with regard to HF operation. At VHF, and higher, we must be concerned about parasitic capacitance C_p and R_s, both of which affect the Q and the upper-frequency limit or cutoff frequency of the diode, F_{co}. The diode cutoff frequency is also affected by L_s.

Fig 1—Electrical equivalent of a VVC diode, showing components of C, L and R (see text).

Types of VVC Diodes

There are three styles of tuning diodes. See Fig 2. The diode at A is the basic single-junction type, with a cathode and an anode. Fig 2B shows a unit that is designed to tune three circuits in an AM broadcast receiver. Three separate VVC diodes are contained in a single case. The tuning diode of Fig 2C features a back-to-back pair of junctions. Single VVC diodes can be connected together as in Fig 2C, if desired.

Diode Q Factor

An important consideration for any resonant circuit is the Q (quality factor). The higher the Q_L (loaded Q), the better the circuit selectivity (sharpness of response). High Q is important to an oscillator because if the Q is too low, the oscillator may not work or it may generate wideband noise. Q is dependent upon,

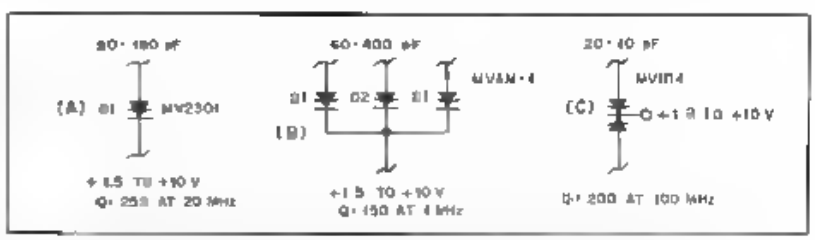

Fig 2—Three types of VVC diodes. A single diode is shown at A. The triple-diode version at B is for tuning three circuits at the same time, such as an RF amplifier, mixer and oscillator. Back-to-back diodes in one package are shown at C.

among other things, the ac resistance of the circuit: the higher the resistance, the lower the Q.

Tuning diodes are rated for Q. This factor varies with the operating voltage and the operating frequency. The Q for a given VVC diode changes considerably as the reverse voltage is varied. The manufacturers' specification sheets include curves showing Q compared with operating voltage and frequency. Check them before selecting a diode for your application.

Performance Trade-offs

Tuning diodes are not perfect! They have some shortcomings that we must take into account as we design circuits using them. They can worsen the frequency drift of a VFO when they are used in place of an air-variable capacitor. This is because all semiconductor junctions change capacitance with changes in junction temperature. Transistor junctions undergo the same changes with respect to temperature. The addition of a VVC diode adds to the short- and long-term drift problems.

Another annoyance with VVC diodes is the fairly high minimum capacitance value. An air-variable capacitor with a range of 50 pF might have a minimum capacitance of, say, 8 pF, whereas a VVC with a 50-pF range can have a minimum capacitance of 25 pF. We must design the tuned circuit to accommodate the high minimum capacitance of the diode. In some cases, this calls for a high C, low L tuned circuit.

The change in diode capacitance is quite nonlinear as the reverse voltage is decreased below approximately 2 V. This means that we should design for operation in a reasonably linear portion of the curve. Fig 3 shows a typical VVC diode voltage/capacitance curve. You can see that there is very little capacitance change from 0.1 to 1 V. If we use this portion of the curve, we will find that our tuning control has minor effect until we reach the 1.5-V region. The range from 2 to 8 V provides a more linear capacitance change, and this is the desired part of the curve. If we allow the voltage to drop below 1.5, a large part of the tuning dial range will be wasted on a 1- or 2-kHz frequency change at, for example, 7 MHz, while the overall frequency change may be 100 kHz in the 2-

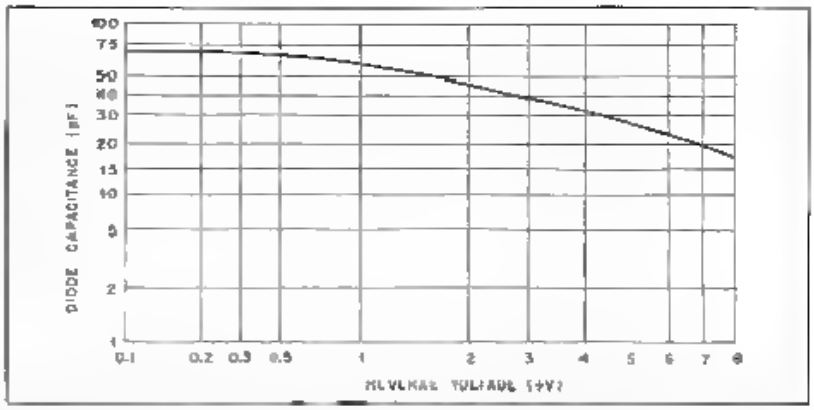

Fig 3—Abbreviated capacitance/reverse voltage curve for an MV2109 tuning diode. Note the flat portion of the curve to the left (see text for precautions about reverse voltages from 0 t to 1.5).

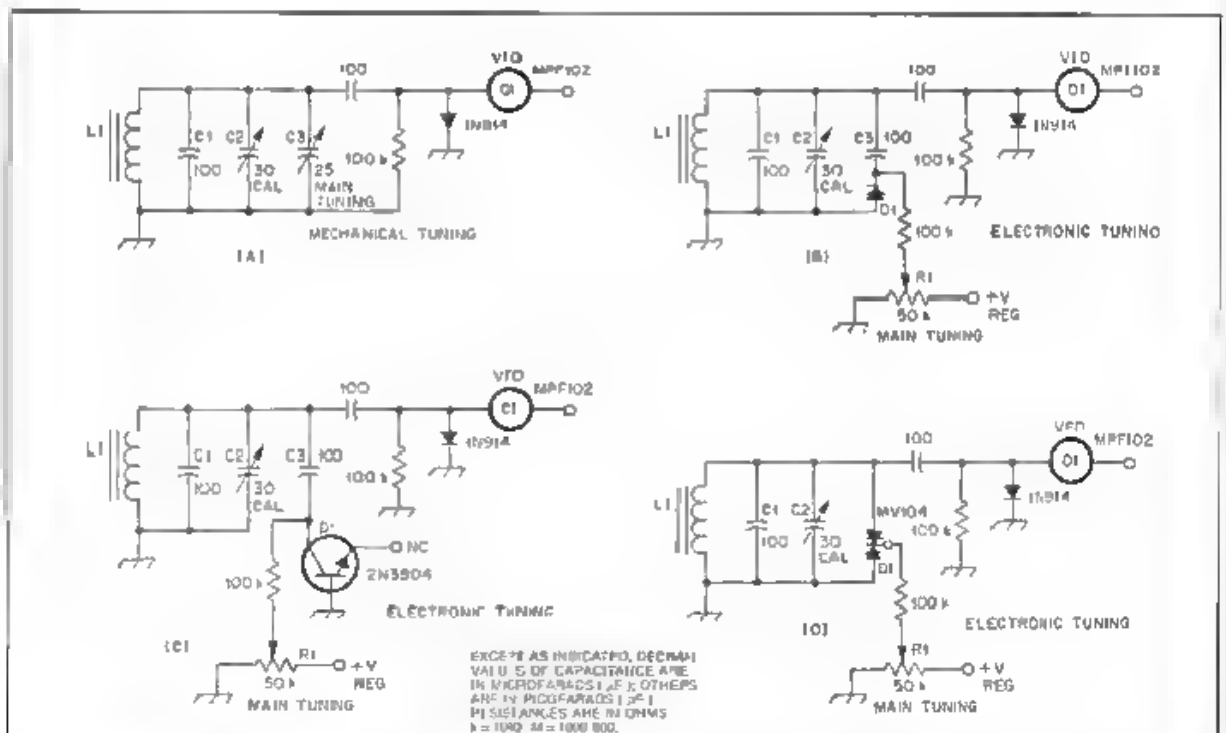

Fig 4—Simplified examples of electronically tuned oscillators. The circuit at A is tuned by a conventional air-variable capacitor. The circuit at B uses a single VVC diode. The example at C shows how to use a bipolar transistor junction as a tuning diode. The circuit at D is preferred, with respect to obtaining a linear oscillator waveform.

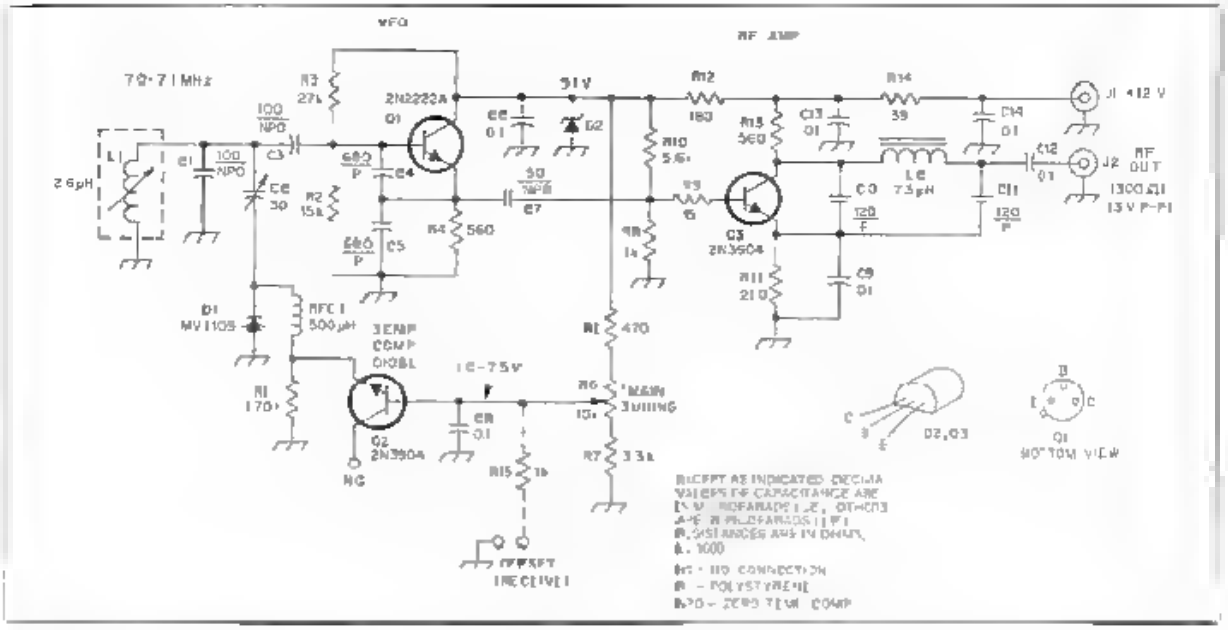

Fig 5—Schematic diagram of a practical VVC-tuned VFO for 40 meters. Unless noted otherwise, fixed-value capacitors are disc ceramic or mylar. Fixed-value resistors are ¼ or ½-W carbon composition, 10% tolerance. Numbered components not listed below are identified for PC-board layout purposes.

C2—30-pF trimmer or ceramic NP0 trimmer (preferred).
D1—Motorola MV2109 VVC diode (see notes 1 and 3).
D2—9.1 V, 400 mW Zener diode.
J1, J2—Single hole mount phono jack.
L1—Slug-tuned inductor, 2.6 µH. Use 16 turns of no. 26 enam. wire on the bobbin of an Amidon Assoc. L-57-6 transformer assembly.
L2—Toroidal inductor, 7.3 µH. Use 36 turns of no. 26 enam. wire on an Amidon T-50-2 core.
R6—Linear taper 10-turn carbon composition potentiometer (see text).
RFC1—Miniature 500-µH RF choke.

to 8-V reverse-voltage range as we adjust the tuning potentiometer.

Most Motorola tuning diodes are rated for +30 V, maximum. I did not extend the curve in Fig 3 beyond 8 volts because the example is used mainly for illustrative purposes. The higher the reverse voltage, the lower the capacitance; but most amateur equipment is designed for 12-V operation. Therefore, we are interested primarily with the portion of the diode curve from 1.0 to 10 V. Regulated reverse voltage is necessary when VVC devices are used for the frequency control of oscillators; this aids frequency stability.

Some Circuit Examples

Fig 4 shows four VFO tuning approaches. Fig 4A illustrates, in abbreviated form, a tuned circuit for a VFO. C3 is an air-variable capacitor used for frequency adjustment. C2 is a trimmer capacitor used for oscillator calibration. The same circuit is seen at Fig 4B, but D1 and its related circuitry replaces C3 of Fig 4A. The values of C1 and L1 may need slight alteration to provide the same frequency coverage that is obtained from the oscillator of Fig 4A. This is because the minimum capacitance of D1 is greater than that of C3 of Fig 4A. R1 is a panel-mounted control used to vary the reverse voltage applied to D1. A resistor may be added between the low end of R1 and ground to prevent the diode voltage from dropping below 1 or 2 V. We can thereby avoid the flat part of the curve of Fig 3.

A transistor junction can be used as a VVC diode as shown in Fig 4C. The capacitance change will differ with the particular transistor used. Some experimentation may be useful.

Fig 4D shows a circuit using a Motorola MV101 VVC diode. Notice that the MV104 features two back-to-back diodes in one case. This arrangement is preferred for better oscillator waveform linearity. Two separate VVCs may be connected together, back-to-back, when we desire to use the method seen at Fig 4D. Similarly, two bipolar transistors (Fig 4C) may be connected back-to-back to permit the arrangement of Fig 4D.

A Practical VVC-Tuned VFO

I chose the circuit of Fig 5 as the local oscillator for a direct-conversion (D-C) receiver I am developing. I want the unit to be small, so I opted for a VVC tuning scheme instead of using a large, expensive air-variable capacitor. I had on hand some surplus Motorola MV2109 Epicap diodes that I purchased from BCD Electro.[1] A capacitance swing of approximately 20 pF is possible in the linear portion of the diode curve. This provides sufficient capacitance change for the circuit of Fig 5, because I am interested in covering only 7.0 to 7.1 MHz. A tuning voltage of 1.6 to 7.5 provides the desired tuning range.

Q2 of Fig 5 is a temperature-compensating device that is connected as a diode. As the ambient temperature changes, so does the resistance of the Q2 diode junction. The small resultant resistance change causes the reverse voltage at D1 to change slightly, thus compensating for changes in the diode junction capacitance that are caused by heat.

R5 and R7 are included to provide the required 1.6 to 7.5 V reverse potential for D1 across R6. You may require different values if the regulated voltage for your oscillator is greater or less than the 9.1 V indicated in Fig 5.

L1 is an adjustable inductor that is wound on the bobbin of an Amidon L-57-6 shielded transformer assembly.[2] The no. 6 (yellow) iron-core material offers good stability in the presence of changing temperatures. No. 2 material (red) has greater permeability (fewer turns needed), but it is less stable than is the no. 6 material. The coil turns are glued in place on the bobbin with a high-quality coil cement, such as General Cement Q Dope.[3] After L1 is tested and adjusted for the desired frequency range, the coil slug should be locked in place by melting a small piece of beeswax or sealing wax onto the end of the coil

[1]Notes appear at end of article.

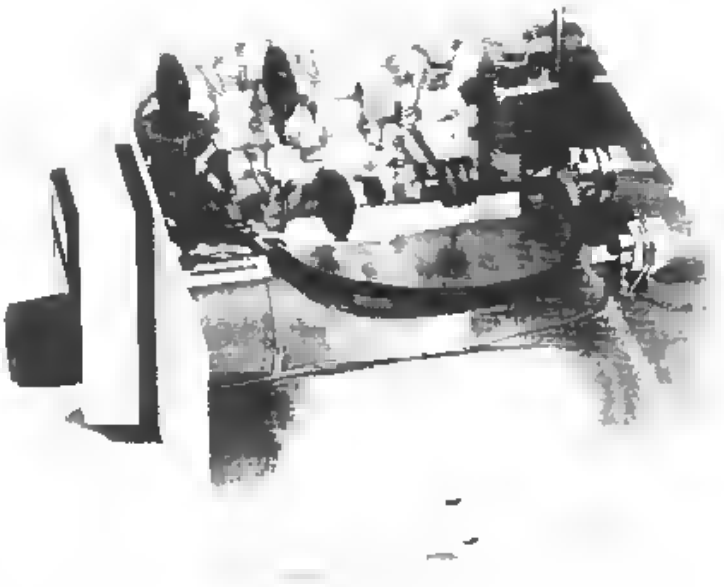

Fig 6—Photographic view of the 40-m VFO in its cabinet with cover removed.

slug. This prevents vibration from moving the slug and changing the oscillator frequency.

C2 is used to set the tuning range of D1. The capacitance of C2 is in series with the capacitance of D1. Therefore, the lower the capacitance of C2, the smaller the frequency spread provided by D1 as R6 is adjusted through its range. In other words, the lower the capacitance at C2, the smaller the effective capacitance change for D1.

NP0 (zero temperature coefficient) ceramic capacitors are used at C1, C3 and C7.[3] Polystyrene capacitors are indicated at C4 and C5, but NP0 units can be used instead. I used polystyrene capacitors because they are quite stable with temperature changes. Also, I did not have a pair of 680-pF NP0 units on hand when I built this VFO. Silver-mica capacitors can be used at C1, C3, C4, C5 and C7 if necessary. You may find that silver-mica units exhibit positive *or* negative drift characteristics, however. Best VFO stability will result if you experiment with these capacitors by trying various units of the same value at each critical circuit point. That is, like-value capacitors of the same brand will often exhibit different drift characteristics with respect to internal heating. For this reason, most homemade highly stable VFOs are practically tailor-made with respect to the final choice of fixed value capacitors in the oscillator circuit.

D2 of Fig 5 regulates the oscillator operating voltage, and ensures a regulated voltage for the D1 tuning circuit. The regulated voltage also stabilizes the forward bias for amplifier Q3, and helps prevent load changes at the oscillator output that would otherwise be reflected by Q3 if the forward bias were allowed to vary.

Q3 is lightly coupled to the emitter of Q1. This also reduces the loading effect of the amplifier. R9 is included as a parasitic suppressor for Q3. If unwanted VHF oscillations are allowed to develop, they will appear at the output of Q3. VHF parasitic oscillations can cause spurious responses in a receiver or transmitter, or cause TVI.

A broadband pi network is used at the output of Q3. It is designed for a Q_L of 3 to ensure a constant output across the VFO tuning range. The network is designed for a 1:1 transformation ratio. R13 sets the collector impedance of Q3. The VFO output impedance (approx 500 ohms) is suitable for interfacing with a class-A bipolar RF amplifier or the 500-ohm input terminal of a mixer IC, such as a CA3028A.

Peak-to-peak output from the circuit of Fig 5 is 3 V across a 470-ohm resistor. This equates to 1.06 V RMS and an output power of 2.4 mW. Greater output power may be obtained by changing R11 to 100 ohms. This provides 5 V P-P or 1.76 V RMS for an output power of 6.6 mW. If greater output power is needed, you may add an RF power amplifier after Q3. A 2N2222A is a good transistor for this purpose. Suitable RF amplifier circuits are presented in *Solid State Design for the Radio Amateur*.[4]

VFO Offset Circuit

Because of the heating of the D1 junction when operating voltage is first applied, you will notice a short-term frequency change of approximately 300 Hz. The VFO settles down and commences its long term drift after about 30 seconds. Because of this, the VFO should remain operating at all times when it is used with a transmitter. R15 of Fig 5 is shown as part of a frequency-offset circuit. During the receive period you may shift the VFO frequency away from the frequency you are listening to by grounding R15. A mechanical or solid-state switch in your TR circuit may be used for this purpose. The amount of frequency offset is determined by the value of R15. If the VFO is used in a D-C transceiver, you may ignore the offset provision—the VFO will be operating at all times.

Drift and Output Waveform

I measured the VFO drift at room temperature (72 °F) with the cover in place on the VFO cabinet. The initial drift took place in a 30-second spurt. Thereafter, the drift was gradual, and stabilization (±2 Hz) was noted after 10 minutes. The long-term drift was measured as 80 Hz. Do not measure your VFO drift for at least an hour after all soldering on the PC board is completed. The VFO module should be mounted in place and enclosed in a cabinet before measurements commence. Even slight stress on the VFO board will cause frequency changes. Solder a 470- or 560-ohm resistor across J2 before doing your drift checks. Set R6 at midrange before starting the test.

I examined the output waveform of Q1 with a 50-MHz Tektronix 453A scope. A clean sine wave was observed and the output amplitude remained constant over the 100-kHz tuning range of the VFO. The filtering action of the Q3 pi network aids in laundering the output waveform.

Practical Considerations

Some type of reduction-gear mechanism is desirable for the VFO of Figs 5 and 6. I used a 10-turn potentiometer and counter dial that I bought at a flea market for $3. Various 10-turn controls and dials are currently manufactured, but the cost may be prohibitive. Check the surplus electronics dealers' catalogs for these mechanisms. You may also use an imported reduction gear drive to control the VVC diode tuning control (R6). If you are adept at making pulleys or metal pulley wheels, try coupling the R6 tuning shaft to the dial-drive shaft with a rubber O ring and two pulleys. A small wheel driving a large pulley wheel will provide a slow tuning rate for the VFO. Some of the small gear drives with readout dials from WW II surplus transmitters, receivers or tuning units can be adapted easily for use as reduction drives for R6.

A good-quality potentiometer is recommended for use at R5. Select a unit that turns smoothly. Industrial-grade controls of the Allen Bradley type (linear taper) operate smoothly, and they will last a long time.

Fig 6 shows the assembled VFO with the cover removed. The unused space in the

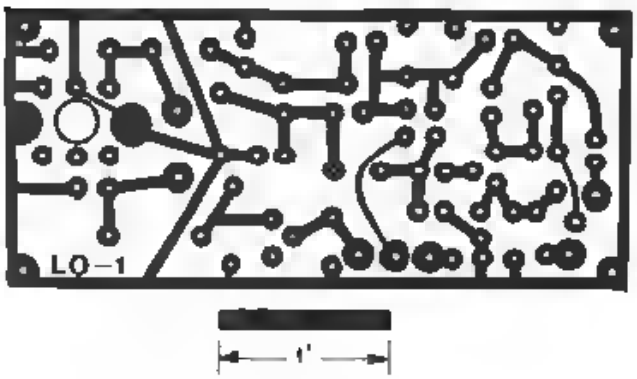

Fig 7—Circuit-board etching pattern for the VVC-tuned 40-m VFO. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil.

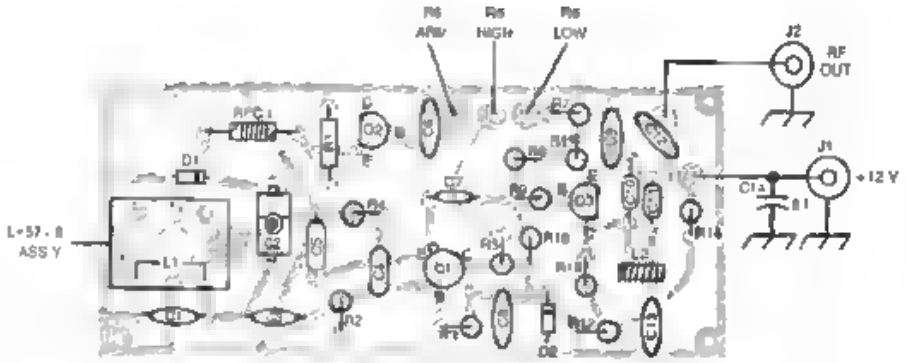

Fig 8—Parts-placement guide for the VVC-tuned 40-m VFO. Parts are placed on the nonfoil side of the board. The shaded area represents an X-ray view of the copper pattern.

cabinet will contain the product detector, active filter and audio amplifier for my 40-meter D-C receiver project. I used a Ten-Tec TP-19 cabinet. It measures (HWD) 2 × 4¼ × 4 inches. As supplied, it is a plain-finish aluminum box. I painted the front and rear panels with gray automotive primer. The cover was also painted with primer, followed by a coat of marine blue gloss enamel. Adhesive-backed plastic feet are affixed to the bottom.

Motorola, Inc, manufactures a variety of VVC tuning diodes.[4] Check with them regarding the availability of data sheets for these diodes. My information came from the *Motorola Semiconductor Library*, Vol 3, series A, 1974 edition.

Fig 7 shows a full-size etching template for the VFO. Fig 8 is a parts-placement guide.

In Summary

I have addressed the subject of VVC tuning diodes in simple terms. The nature of these diodes is considerably more complex than this article indicates. However, you should now have sufficient knowledge to permit plenty of experimentation and practical satisfaction. Application notes from the companies that manufacture VVC tuning diodes will give you greater insight into the performance characteristics of these devices. If nothing more, you can save money by using tuning diodes, and your equipment will be much smaller than when using mechanical tuning capacitors.

Notes

[1]Deleted

[2]Amidon Assoc, Inc, 12033 Otsego St, N Hollywood, CA 91607. Catalog available.

[3]NPO capacitors, silver-mica capacitors and MV2109 VVC diodes (also other VVC diodes) are available from Circuit Specialists, PO Box 3047, Scottsdale, AZ 85257, tel 602-966-0764. Catalog available.

[4]W Heyward and D DeMaw, *Solid State Design for the Radio Amateur*, 2nd printing (Newington: ARRL, 1986)

[5]Motorola Semiconductor Products, Inc, Technical Information Center, PO Box 20924, Phoenix, AZ 85036

For updated supplier addresses see ARRL Parts Suppliers List in Chapter 2.

From January 1989 QST, p 31

A VFO with Bandspread and Bandset

Eliminate expensive vernier drives and dials with an old technique—bandspread and bandset tuning!

By Doug DeMaw, W1FB
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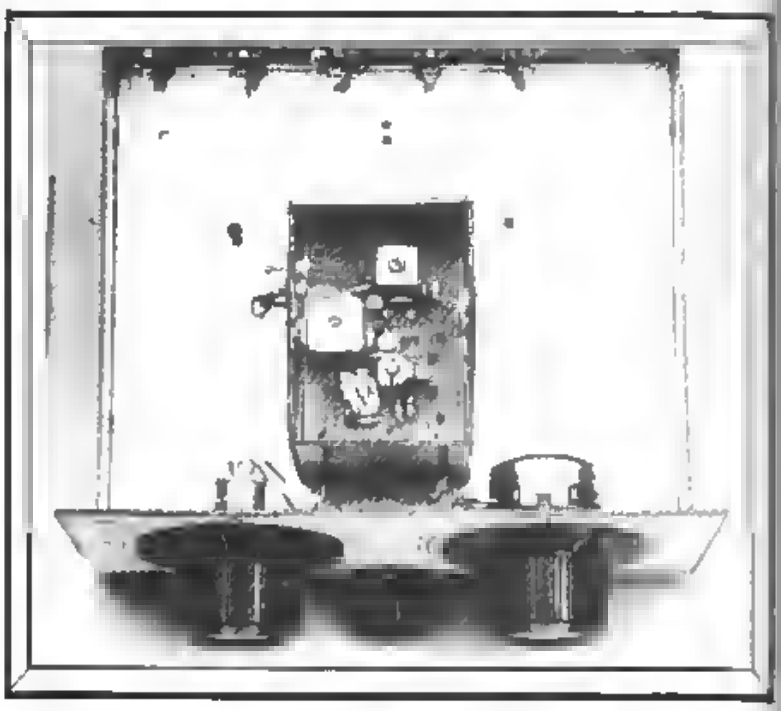


Are you old enough to recall those days when we amateurs had receivers that had two readout dials? One was a bandset dial (coarse tuning) and the other was for bandspread (fine tuning). When I compare that method to modern digital-readout techniques, I wander how we managed to get on frequency; the resolution of the dials was primitive by today's standards! The bandset dial was calibrated in megahertz and the bandspread dial indicated kilohertz. The tuning increments for the bandspread dial were in 5- or 10-kHz steps, depending on the model of the receiver.

We may apply that old technique to modern circuits. Reasonable readout accuracy is possible with the method discussed in this article. The trick is to make both dials read kilohertz, rather than megahertz and kilohertz. The circuit described here is meant to be an inspiration toward a design of your own. It serves as a model for a starting point, with a circuit board pattern offered if you wish to experiment. My circuit values are for use in a 6.572- to 6.872 MHz VFO. This VFO serves as the local oscillator for a homemade 80-meter CW receiver that uses a 3072-kHz IF and a crystal filter made from low-cost computer crystals. I plan to describe the entire receiver in a subsequent article.

Circuit Features

Please refer to Fig 1, which shows the circuit for my experimental VFO. You will note that I use electronic tuning. D2 and D3 are VVC (voltage variable capacitance) diodes. They are also called varactors or tuning diodes. As the reverse bias (positive voltage) is varied at the diode cathode, there is a significant change in the junction capacitance of the diode. This enables us to change the VFO frequency, as would be the situation if we replaced D2 and D3 with mechanical tuning capacitors. The advantage of using the diodes is that we can use standard carbon-composition controls (R2 and R7 of Fig 1) for tuning the VFO. This provides a compact VFO module, should that be our objective.

D2 functions as the bandset tuning diode, while D3 is used for the bandspread function. Each diode has a trimmer capacitor (C3 and C4) between it and L2. The trimmers are set to control the tuning range of each VVC diode.

All is not "milk and honey" when we use tuning diodes in VFOs. Although the diodes offer some advantages over air-variable capacitors, they are not as frequency stable as mechanical tuning devices. The more semiconductor junctions we add to an oscillator circuit, the greater the opportunity for frequency drift—particularly short-term drift (first five minutes of warm-up). This is because the transistor and diode junctions must come up to operating temperature as current flows through them. This involves both RF and direct currents. The stability of the VFO in Fig 1 is adequate for many amateur needs, such as simple receivers and signal generators. Short-term drift is on the order of 1.5 kHz from a cold start to the period when long-term drift commences. Long-term drift occurs for 15 or 20 minutes, and it amounts to a range of 200-300 Hz. Thereafter, the frequency creeps up and down over a range of 5-10 Hz at room temperature. In other words, the circuit in Fig 1 represents a good VFO, but not a spectacular one. It is on par with what I expected when using two VVC diodes.

Remainder of the VFO Circuit

Q1 of Fig 1 is a 2N4416 JFET. This device surpasses the performance of the generic MPF102 family of transistors. It has a better pinchoff characteristic than does the MPF102 and similar devices. This means that greater output is possible at a given operating voltage, compared to an MPF102. Oscillator feedback is by way of the Q1 source and L1. This link has ¼ the number of turns used on L2, which is pretty standard for a feedback winding. The two coils are wound on an Amidon L-57-6

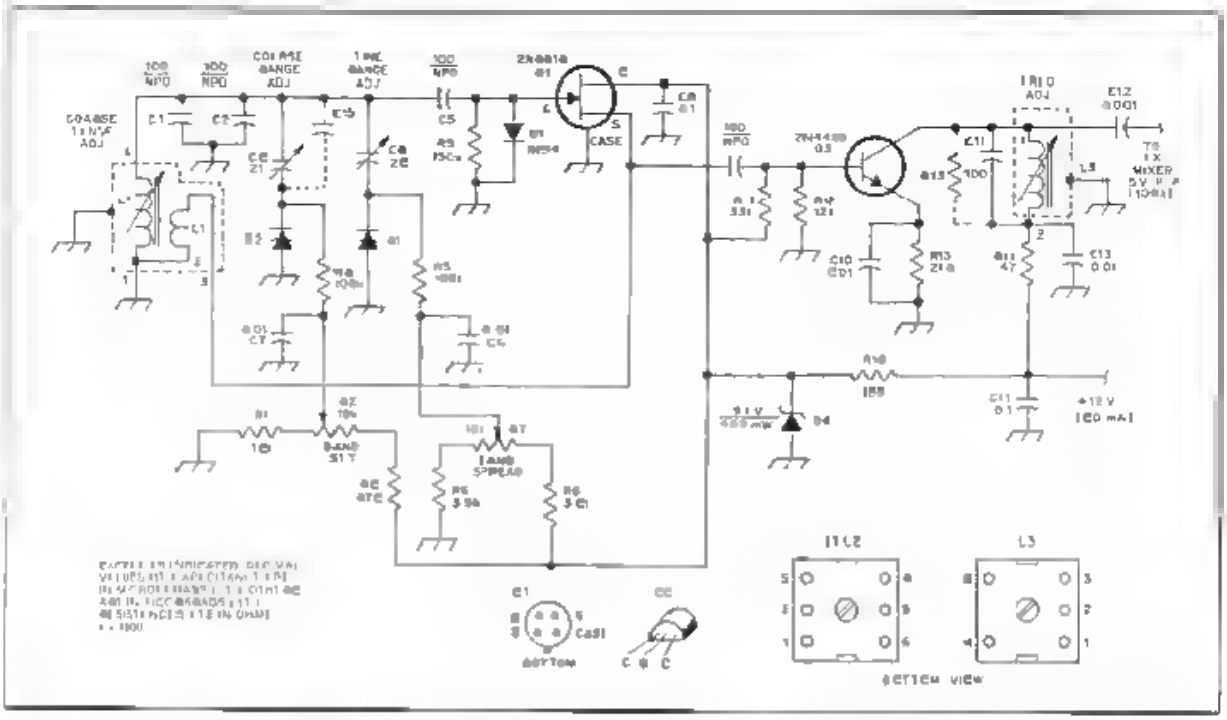


Fig 1—Schematic diagram of the VVC-tuned VFO. Fixed-value capacitors are disc ceramic, 50- or 100-V rating. Fixed value resistors are ¼-W carbon composition. NPO notations are for temperature-stable disc capacitors (zero temperature coefficient).

C3, C5—25-pF NPO miniature ceramic trimmer or E. F. Johnson 9-5 miniature air-variable trimmers.
C15—See text.
D1—Silicon switching diode, type 1N914 or equiv.
D2, D3—Motorola MV2109 tuning diode or equiv. 30-70 pF typical range. Available from All-Electronics Corp, Van Nuys, CA 91406.
D4—9.1-V, 400-mW Zener diode.
L1—4 turns of no. 30 enam. on Liiz wire over grounded end of L2. Use Amidon Assoc, Inc. L-57-6 shielded assembly.
L2—16 turns of no. 30 enam. on Liiz wire on L-57-6 bobbin. Use Q-Dope to secure windings (see text).
L3—24 turns of no. 30 enam. on Liiz wire on the form of an Amidon Assoc. L-43-2 shielded assembly. Turns must be scramble-wound to fit on form.
R2, R7—10-kΩ linear-taper carbon-composition control (see text).
R15—See text.

transformer assembly. The no. 6 (yellow) powdered-iron core material is best for VFO service. It is more temperature stable than the other core materials.

NP0 temperature-stable capacitors (C1, C2, C5 and C9) are used to aid the stability. D1, from the Q1 gate to ground, stabilizes the bias on Q1 and limits the device transconductance on sine wave peaks. This helps to keep the junction capacitance fairly constant—an aid to stability. A further enhancement to stability is provided by Zener diode D4. It regulates the operating voltage for D2, D3, Q1 and the base of Q2.

Buffer amplifier Q2 is used to boost the RF output of the oscillator chain to 5 V P-P. The output is designed to look into a 100-kΩ load, which may be gate no. 2 of a dual-gate MOSFET mixer. R15 may be added (3.3 kΩ to 10 kΩ) across L2 (dashed lines in Fig 1) to broaden the response of L2. This will reduce the RF output somewhat.

You may use a lower value of capacitance at C9 if you require lower output from Q2. The smaller the C9 value, the greater the overall VFO stability. In a like manner, the lower the C5 value, the better the stability. C5 needs to be of a large enough value to allow Q1 to oscillate. The Q of the oscillator tank and the specific transconductance of Q1 are determining factors when selecting the C5 value in a VFO of this general type. C5 values as low as 5 pF are usable, especially when L2 has a high value of Q (100 or greater).

Circuit Variations

If you desire greater frequency stability than I mentioned earlier, replace D2 and D3 with small air-variable capacitors. You may use a 100-pF unit in place of D2. The bandspread tuning can then be done with a 15- or 20-pF variable. This calls for the deletion of the VVC diode components, R1 through R8, plus C6 and C7, and of course, D2 and D3.

C15 of Fig 1 is shown in dashed lines. You may add a capacitor at this circuit point if you wish to increase the tuning range of the bandset control. Experiment with the C15 value to obtain the range you need.

Construction In General

Use a single-sided PC board for this project. Double sided board material increases the VFO drift, owing to the formation of unwanted low-stability capacitance between the PC foil and the ground-plane side of the board. Try to use high quality glass-epoxy board material. Phenolic PC boards are not suitable for VFOs.

I enclosed my VFO in a homemade box, as shown in the title photo. The box is made from pieces of PC board that have been soldered together. The cover, removed for the photograph, is a U-shaped piece of aluminum. The cover is affixed to the box by means of two no. 4-40 screws. I soldered two 4-40 × ¼-inch nuts on the inside of the box to accommodate the two screws. I used two surplus Teflon push-in feed-through terminals to route the +12 V to the circuit, and to bring the RF output from

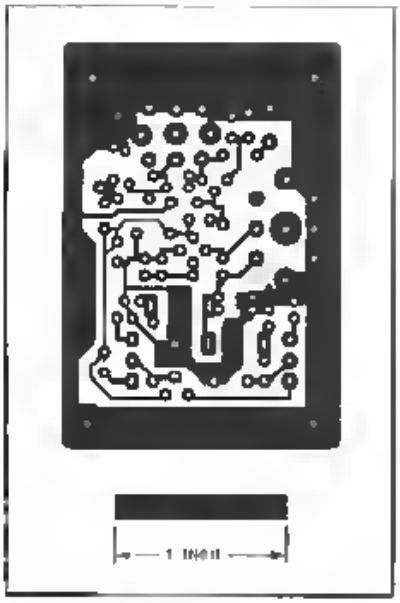


Fig 2—Circuit-board etching pattern for the VVC VFO. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil.

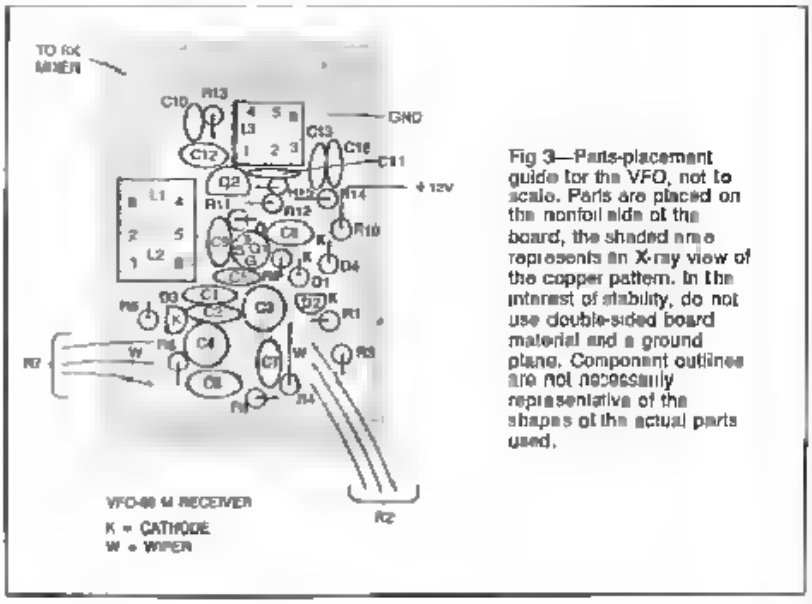


Fig 3—Parts-placement guide for the VFO, not to scale. Parts are placed on the nonfoil side of the board, the shaded area represents an X-ray view of the copper pattern. In the interest of stability, do not use double-sided board material and a ground plane. Component outlines are not necessarily representative of the shapes of the actual parts used.

the box. Two no. 6 spade bolts secure the VFO box to the mainframe assembly that will later contain the remainder of my 80-meter receiver. The hookup-wire cables for tuning controls R2 and R7 are brought from the VFO box through ¼-inch holes in the box wall. The VFO module measures (HWD) 2 × 2 × 2½ inches.

A scale etching pattern for the VFO PC board is provided in Fig 2. A parts-placement guide is shown in Fig 3.[1]

The dial-calibration plate for my VFO is homemade, visible on edge in the title-page photograph. I drew the circles with a ballpoint pen and compass. I use knobs with large skirts (2 inches OD), bought at a flea market. If you can't locate a pair of large knobs with skirts, you may use standard-size knobs and metal or plastic dial skirts with them. The skirts may be attached to the knobs by means of epoxy cement or small screws.

After I made the dial plate I photocopied it. The copy was used for dial calibration with a pencil. I measured the VFO output with a frequency counter. My VFO is set for 50 kHz of tuning range with the bandset control. The bandspread covers only 10 kHz. Midrange on the bandspread control is marked zero. To the left of zero I calibrated the dial with minus kHz marks. Plus-kHz marks are to the right of zero. After plotting the calibration scales I made marks between the two rings of each dial face, then typed the frequencies alongside the marks. Rubber cement is used to affix the dial plate to the front panel.

Try to obtain commercial-grade controls for R2 and R7 of Fig 1, such as Allen-Bradley units. They will last longer than imported controls, and will be less prone to resistive instability from shock and vibration. Check the surplus catalogs for these controls.

Checkout and Operation

You will need to adjust the slug in L2, along with the settings for C3 and C4. First, determine how much frequency range you want to cover with the bandset control. Adjust C3 and L2 so R2 provides the desired range. Next, adjust C4 to yield 10 kHz of tuning range for R7. This will cause some interaction with the settings of C3 and L2. Repeat those adjustments to obtain the desired tuning range for R2.

Next, terminate C12 with a 100-kΩ resistor. Connect a scope or RF probe from the output side of C12 to ground. Adjust the slug in L3 for maximum RF output voltage.

L1 and L2 should be coated with GC polystyrene Q-Dope after they are wound on the L-57-6 bobbin. Allow at least 48 hours for the coil to dry before you check the stability of your VFO. Q-Dope is available by mail from Small Parts Center.[2] Do not attempt drift tests if you have recently soldered connections on the VFO PC board. Allow an hour after all soldering is completed before you commence your drift run. Keep the module away from desk or bench lamps and enclose the VFO PC board in its box to prevent air currents from reaching the critical components. Terminate the VFO output with a 100-kΩ resistor and attach a frequency counter to the VFO output through a 27- or 33-pF capacitor. Apply the VFO operating voltage and log the initial frequency. Monitor the frequency change until the drift is only 1 or 2 Hz per count. Observe the frequency change until it stabilizes. This will be noted when the frequency shifts up and down by a few hertz in a random manner. Dial calibration (discussed earlier) should be done *after* the short-term drift has occurred. This should take place within five minutes after turn-on.

Closing Comments

I want to stress that this is an "idea" article rather than a project for duplication. The main thought here is that you can capitalize on the old technique of using a bandset and bandspread setup in order to avoid the high cost of vernier mechanisms. Tuning diodes are discussed in the interest of equipment miniaturization and reduced cost.

This VFO is not recommended for use with transmitters unless one or more additional stages of buffering are used. A single buffer-amplifier does not provide the load isolation that is necessary between the VFO and a transmitter. It is adequate, however, for connection to a mixer that presents a relatively constant load impedance.

You should have no difficulty in tailoring this circuit to other frequencies. All that is necessary is to change the inductance of L2 and L3, along with appropriate modifications for the values of C1, C2, C5, C9 and C11. I'm sure you will have fun experimenting with this circuit, and you can learn by doing!

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

Far Circuits (N9ATW), 18N640 Field Court, Dundee, IL 60118, tel 312-426-2438 evenings

[2]Small Parts Center, 6818 Moose Drive, Lansing, MI 48911, tel 517-882-6447. Catalog available.

From September 1978 QST, p 15:

Meet the Remarkable but Little-Known Vackar VFO!

Searching for a VFO with Rock of Gibraltar stability? End your band-edge worries with this self-contained unit. For the serious-minded cw operator, the chirp-free operation and undetectable frequency drift make this VFO a natural!

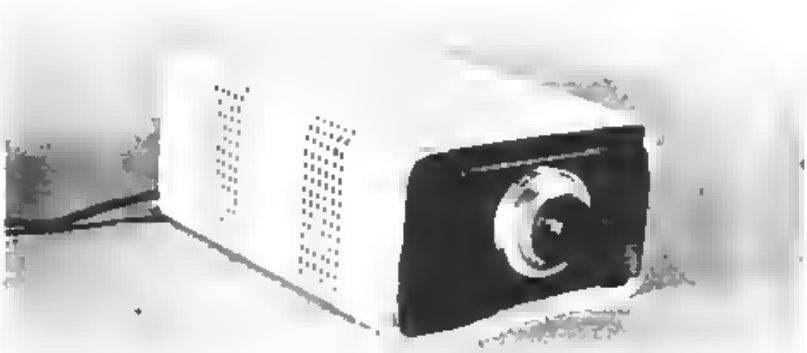
By Floyd E Carter,* K6BSU

The dedicated cw operator must make severe demands of his station equipment. He knows that an elusive DX station amateur cannot be asked to tolerate a signal which drifts through the passband of his receiver or one which has keying chirp. For the cw man, his fist and the note of his transmitter form his "voice" to distant stations. Modern electronic keyers have made machine-like keying an inexpensive reality. Couple a keyer with a fine-quality VFO, and the DX station operator just cannot refuse to QSO.

In designing this heterodyne VFO, the goal was to produce a keyed oscillator with undetectable chirp or frequency drift. Keying of a conventional VFO invariably produces some instability because the starting and stopping of an oscillator upsets the fine balance of dc and rf conditions within the circuit, and with each key-down transition oscillation equilibrium must be reached. During this transient period, the oscillation frequency generally changes, resulting in chirp. Keying of a subsequent buffer stage following a free-running VFO generally allows a small portion of the VFO output to reach the receiver during key-up conditions if the station is set up for full-break-in cw. VFO shielding only reduces the feedthrough, but this may not be adequate for very sensitive station receivers.

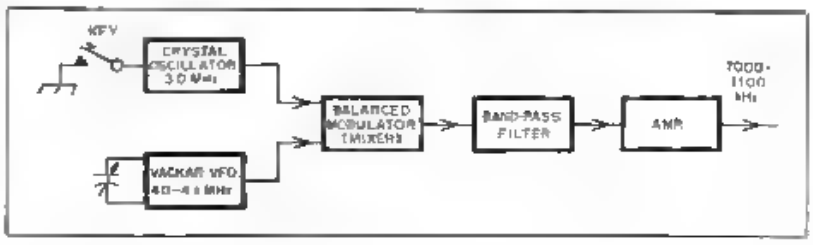
Heterodyne frequency generation eliminates all these problems because the VFO operates continuously on a non-harmonically related frequency which is converted to the operating frequency in a mixer or balanced modulator. Both the keyed crystal oscillator and the VFO operate far from the receiver frequency. Therefore, even though the VFO is not keyed, no harmonic of the oscillator will reach the receiver. Fig 1 shows the block diagram of the heterodyne process, with frequency values applicable to this VFO

*2029 Crist Dr Los Altos, CA 94022



The Vackar oscillator VFO enclosed in an attractive contemporary styled cabinet. Below is an inside view showing rather high component density. The U3 output amplifier is on a separate board next to the transformer.

Fig. 1 — Simplified block diagram of the heterodyne VFO



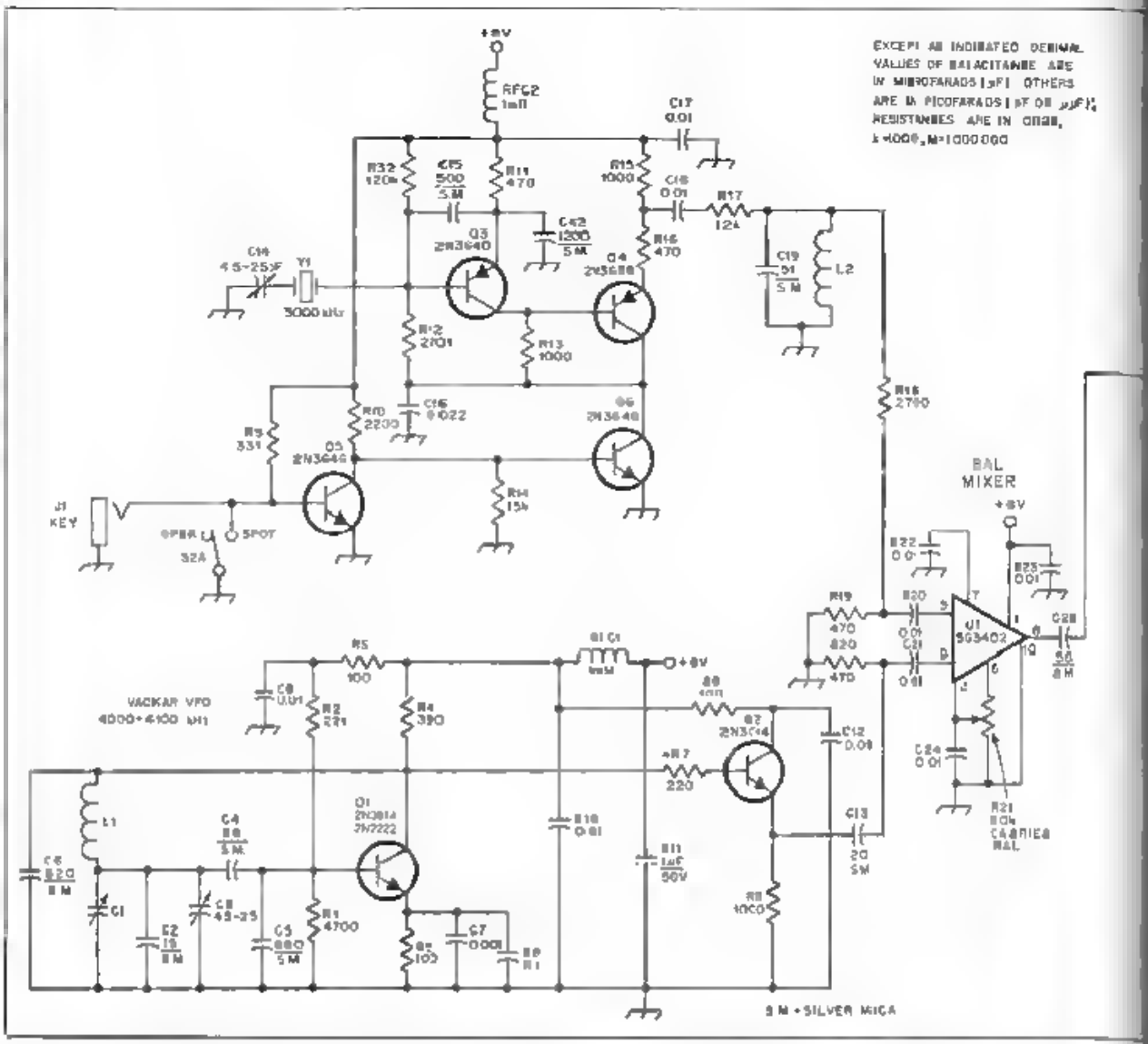


Fig 2 — Schematic diagram of the heterodyne-oscillator VFO using the Vackar circuit. All resistors are ¼-watt, five-percent tolerance. U1 is a proprietary product manufactured by Silicon General, Inc., 7382 Bolsa Ave., Westminster, CA 92683. The toroid core for L2, Ferroxcube no. 1041T060/4C4, is produced by the Ferroxcube Corp., Mt. Marion Rd., Saugerties, NY 12477. (For the convenience of builders who are unable to locate small toroids, the author has available a limited supply.) For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

A normal mixer or unbalanced modulator output contains four prominent frequency components — the two input frequencies, their sum, and their difference. Either the sum or the difference may be used as an output by selecting the desired frequency in a band-pass filter. The balanced mixer is a more sophisticated refinement of the basic mixer circuit, because the two input frequencies are eliminated in the mixing process so that the output contains only the sum and difference frequencies. Consequently, subsequent filtering is made easier.

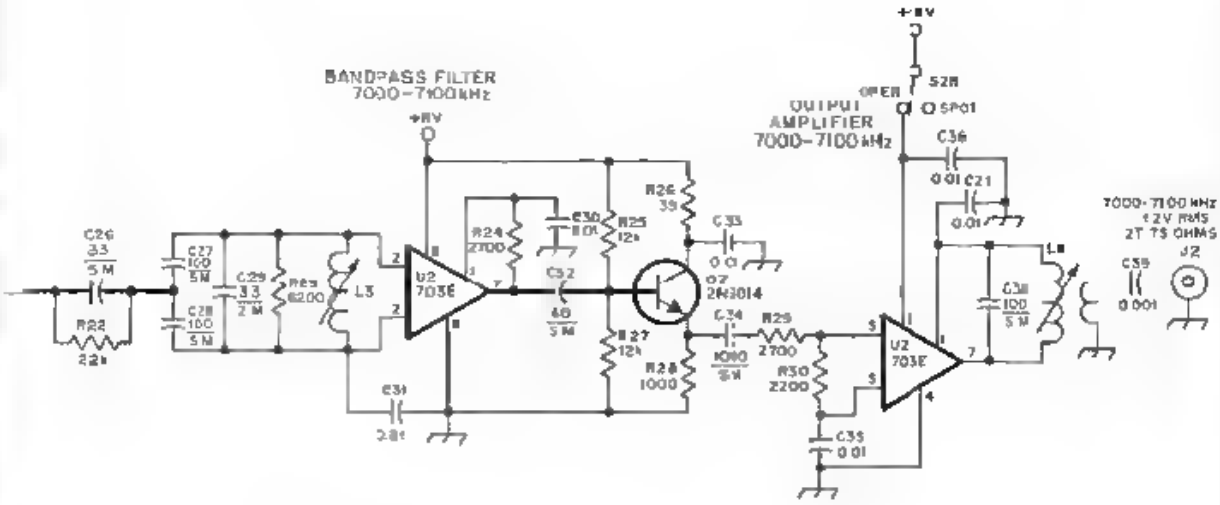
The VFO circuit used in the heterodyne VFO was first described by Vackar[1] in

[1]Notes appear at end of article.

1949. This circuit formed the basis for further research by Clapp, resulting in his classic article published in 1954.[1] The Vackar circuit closely resembles the Clapp circuit except for the method of feedback. The Vackar is series tuned like the Clapp, but the tank circuit as well as the transistor are shunted by unusually low reactances which reduce the effects of the transistor reactances. Further refinements of the Vackar circuit were described in 1968 by Jordan,[2] who provides design criteria for use at any frequency.

Construction

The photographs suggest one possible layout. For ease of modification and experimentation, the prototype was built in separate modular form equipped with connectors. Only a few precautions must be kept in mind when designing a layout. First, as with any VFO, mechanical stability is essential. An aluminum extrusion was used as a base for the oscillator. The tank components were bolted to this extrusion and the remainder of the circuit is contained on a glass-epoxy board bolted to one lip of the extrusion. Heavy solid wire is used to interconnect the tank circuit components to prevent changes in stray circuit capacitance from shock or vibration. The integrated circuits have much higher bandwidths than required, and are capable of oscillations at vhf.



C1 — Variable capacitor, approximately 2 pF (1 rotor and 1 stator).
C3, C14 — 4.5-28 pF variable capacitor, CRL no. 825-AZ.
C7, C38 — Fixed capacitor, 0.001 µF, CR, no. CE102.
C8 — Fixed capacitor, 0.1 µF, CRL no. DDA104.
C9, C10, C12, C17, C18, C20-C24 incl., C30, C31, C33, C35, C36, C37 — Fixed capacitor, 0.01 µF, CRL no. CK103.
C19 — Mylar fixed capacitor, 0.022 µF, CDE no. 1622.
C40 — Fixed capacitor, 1000 µF, 25 V dc, CDE no. HWM 1000-25 (Fig. 3).
C41 — Fixed capacitor, 500 µF, 10 V dc, CDE no. HWM 500-15 (Fig. 3).
D1 — Silicon voltage regulator diode, 8.2 V, 400 mW, Texas Instrument no. 1N756A or equiv. (Fig. 3).
J1 — 1/4-inch phone jack, Switchcraft no. 11.
J2 — Chassis mount jack, Switchcraft no. 3505F.
L1 — 18 µH, 31 turns No. 22, enameled copper wire, 7/8 inch long, 1 inch diameter. Ceramic form, National no. XR-50.
L2 — Toroid core, Ferroxcube no. 10417060r 4C4, approximately 50 turns no. 28 enameled copper wire.
L3 — Miller no. 42A000CB1-2, 28 turns no. 24 enameled copper wire.
L4 — Miller no. 40A000CB1-2, primary 26 turns no. 28 enameled copper wire, 3/8 inch long, secondary 12 turns no. 28 enameled copper wire.
Q1 — Npn silicon annular transistor, type 2N918 or equiv.
Q2, Q7 — Npn silicon annular transistor, type 2N3014 or equiv.
Q3 — Pnp silicon low-power transistor, type 2N3640 or equiv.
Q4 — Pnp silicon high-current switching transistor, type 2N3638 or equiv.
Q5, Q6 — Npn silicon low-power transistor, National Semiconductor type 2N3646 or equiv.
Q8 — Npn silicon annular transistor, type 2N697 (Fig. 3).
S1 — SPDT toggle switch, Alco no. MST-105D.
S2 — DPDT toggle switch, Alco MST-205N.
U1 — Variable gain, wideband amplifier/multiplier, Silicon General no. SG3402.
U2, U3 — Linear IC, monolithic rf i-f amplifier, Fairchild no. 703E.
U4 — Silicon miniature diode assembly, Motorola MDA 650-2 or equiv. (Fig. 3).
Y1 — Oscillator crystal, 3000 kHz. Sources listed in QST advertisements.

Therefore the bypass capacitors should be mounted close to the IC with short leads. The planetary ball reduction gear couples the tuning capacitor to the tuning knob. This is not an ideal setup for it is not possible to calibrate the dial because the ball drive slips at the end of travel. However, accurate calibration of a VFO is not a great advantage, inasmuch as crystal band-edge markers are required if one is going to operate within striking distance of a pink slip.

Test and Adjustment

The only tuned circuit which is not adjustable is the 3-MHz band-pass filter consisting of L2 and C19. This should be resonated with a grid-dip meter after first overwinding the toroid core and removing turns one at a time until the circuit resonates. This circuit removes harmonics from the crystal oscillator and helps to reduce spurious inputs to the balanced modulator.

With the VFO operating and keyed, the output of U1 should be monitored while adjusting R21, the carrier-balance potentiometer, for a null at both 3 MHz and 4 MHz. The null should occur simultaneously. Next, monitor the output of J2 through a length of coaxial cable terminated in the transmitter. The cable is necessary because the cable capacitance is reflected back into the circuit for L4 and C38 and forms part of the total tuning capacitance. Adjust L3 and L4 for maximum drive to the transmitter. While rapidly keying the crystal oscillator, adjust C14 for the best starting characteristics. Finally, C1 is adjusted to cover the spread of 4.0 to 4.1 MHz. Adjustment is made with C3 and by bending the plates of C1 for the desired delta C for full rotation.

If a spectrum analyzer is available, the optimum tuning may be quickly reached for maximum rejection of unwanted frequency components. The prototype circuit had all unwanted components down by at least 40 dB. With key up, the VFO feedthrough at 4 MHz was down 30 dB. This level is not detectable with the station receiver and tuned circuits in the driven transmitter will reject these components.

With S2 in the SPOT position, power is removed from the output buffer amplifier and the crystal oscillator is keyed. This

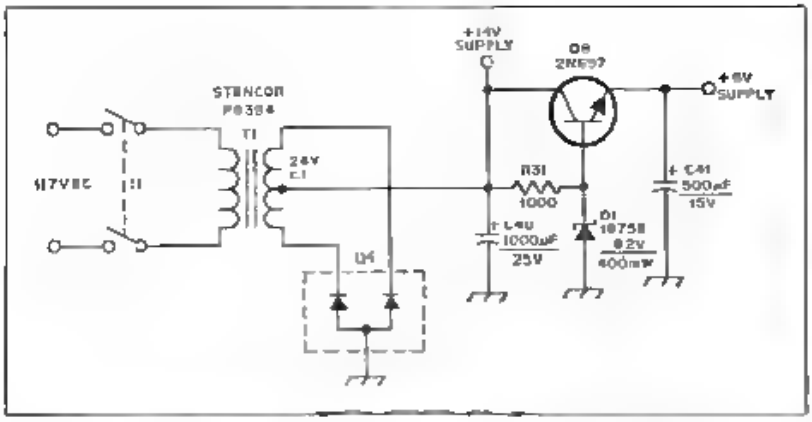
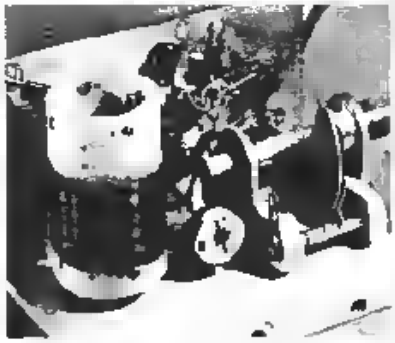
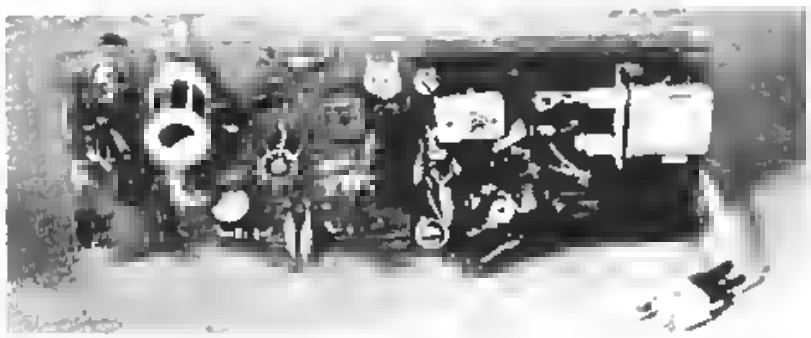


Fig 3 — Power supply for the heterodyne VFO. Miniature diode assembly U4 is Motorola part no MDA950-2 or equiv.



The Vackar oscillator circuit is constructed on a heavy extrusion. Large bus wire interconnects fused circuit components. L1 is wound on a ceramic form and coated with epoxy resin. C1 is a heavy-duty two bearing capacitor reduced to one rotor and one stator plate.



Crystal oscillator and balanced mixer board. The oscillator is a highly modified International AOI assembly. The small toroid coil on the oscillator board is L2. The balanced mixer (10-lead IC) is on the main board. U2 and Q6 are at far left. As is typical with developmental circuits, the board shows evidence of modifications.

generates a weak signal which can be monitored in the station receiver for frequency spotting. In the OPERATE position, control is transferred to the keyer. Any commercial keyer with an open-collector, current-sinking output will work with this VFO. If there is doubt in one's mind about this feature of a particular keyer, the schematic diagram of the keyer should be examined, or the manufacturer should be consulted. Of course, a relay output will also work with the VFO.

The normal output of the heterodyne VFO is about 20 mW into a load of 75 ohms. The driven transmitter operates straight through on 40 meters for outputs of 7.0-7.1 MHz. Using the driven transmitter as a multiplier, 20-meter output from 14.0-14.2 or 10-meter output from 28.0-28.4 MHz is available. The driven transmitter must also be provided with fixed bias to prevent excessive dissipation in the final amplifier under key-up conditions. For transmitters with cathode or emitter keying, fixed bias should be added to cut off the final amplifier during key-up conditions.

The heterodyne VFO has been in use with a Viking II transmitter with the tension set up for full break-in cw operation. It is the only VFO I have ever used where operation very close to the band edges in the Extra Class portion is possible without constant nervous strain from wondering just where the transmitted frequency will end up after a long QSO.

References

1. Vackar, "LC Oscillators and Their Frequency Stability," *Tesla Technical Reports* (Czechoslovakia), Dec., 1949.
2. Clapp, "Frequency Stable LC Oscillators," *Proc. IRE*, Aug., 1954, pp. 1295-1300.
3. Jordan, "The Vackar VFO: A Design To Try," *Electronic Engineer*, Feb., 1968.

From April 1989 QST, p. 38:

Adjusting the Power Output of JFET VFOs

☐ The output of a JFET VFO is determined largely by the device standing current—the JFET's drain current with dc bias applied and ac feedback removed. In many VFO designs, this is equivalent to I_{DSS}—the zero-gate-voltage drain current. Generally, the relationship between I_{DSS} and oscillator output is simple: The higher the device I_{DSS}, the greater the VFO output.

According to the *Motorola Small-Signal Transistor Data* book, I_{DSS} for the popular MPF102 can fall anywhere within the wide range of 2 to 20 mA. This wide I_{DSS} specification explains why some VFO builders have good luck with the MPF102 and others build MPF102 VFOs that deliver less output than that claimed for the circuit involved. The "premium" 2N4416 has an I_{DSS} range of 5 to 15 mA, making the '44 6 generally better than the MPF102 if you want more power output. The best commonly available JFET for lots of VFO output is the 2N5486, which has an I_{DSS} range of 8 to 20 mA.

It's important to keep another rule of thumb in mind: Oscillator frequency stability generally *decreases* as power output increases. If you're willing to sacrifice VFO output for greater frequency stability, the 2N5484 (I_{DSS} of 1 .o 5 mA) and 2N5485 (I_{DSS} of 4 to 10 mA) are good choices.

By the way, the resistance of the JFET channel is a good relative indicator of device I_{DSS}. With this in mind, you can grade your JFETs for VFO power output merely by measuring their channel resistance (source to drain) with a DMM. (Caution: The measuring instrument you use must not apply a destructively large current to the device under test.) Generally, the lower the channel resistance of a given device, the more power output it will furnish as a VFO.—*Zack Lau, KH6CP, ARRL Laboratory Engineer*

From April 1979 QST, p 18:

Putting the Boots to Your HW-8 QRP Transceiver

Basic Amateur Radio: A signal increase of 9 dB for your QRP rig can turn marginal QSOs into solid ones! This amplifier provides 80- through 15-meter signal increases with only 1 watt of drive. Add these "boots" to your HW-8 and improve your QRP DX score.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther, MI 49656

An article describing a single-ended plug-in amplifier for the HW-7 series QRP transceiver left much to be desired before it became *QST* readers.' ARRL hq was hit with a slash of letters requesting a band-switching style of amplifier which included the 80-meter band along with the 40-, 20- and 15-meter bands. Coverage on 80 meters was suspected by the appearance of Heath's newer QRP box, the HW-8. The additional cost of single-band amplifiers over a band-switching unit was objectionable to some, and eighty so. Moreover, some builders reported problems with amplifier instability when they assembled the "Slippers" unit. The amplifier described here is aimed at HW-8 owners in particular. However, it can be used with any QRP transmitter if the latter has the output attenuated so that a maximum of 1 watt reaches the power amplifier input. It is a simple matter to install an appropriate T or pi type of resistive attenuator at the amplifier input when more driving power than is necessary appears at the output of the QRP transmitter.

This amplifier operates from a 12- to 14-volt dc supply. Maximum current drain is less than 2 amperes. A spectral analysis of the amplifier output indicated that its amplifier with the FCC requirement that all spurious energy be 40 dB or greater below peak carrier value. The spurious components are −45 dB or better on each of the bands covered by this circuit. A test of an early prototype were observed at levels well below −40 dB. They are products generated within the HW-8 and do not originate in the amplifier described here.

Notes appear at end of article

Circuit Description

A pair of RCA 40977 stud-mount power transistors are shown in the circuit of Fig. 1. These are actually vhf devices and are used primarily because they were on hand at the time this circuit was developed. Later, RCA dropped this part from the line. Transistors with similar characteristics for hf-band operation may be used in place of the 40977, notably the Motorola 2N5642 which is an exact replacement. The specifications for the 40977 are 11 dB gain (approximate) at 118 MHz. Look for a substitute which has similar gain at 21 or 30 MHz. Maximum power dissipation is 25 watts. Power input is 0.5 watt (approximate) for 6 watts minimum output. Collector supply voltage is 12.5 nominal. Continuous collector current rating (maximum) is 5 A. Collector efficiency is 55 percent. The builder should not be afraid to experiment with other types of power transistors, especially if they can be obtained inexpensively as surplus from a reliable dealer.

This circuit operates broadband in the Class C mode. This technique simplifies band switching and lowers the cost. To ensure unconditional amplifier stability it is necessary to use shunt feedback from collector to base (R1, R2, C1, C2, L1 and L2). Broadbanding and stabilization of this type always results in a power trade-off. In a similar circuit which used no feedback, the amplifier output could be as great as 15 watts, safely, even though the 40977s are rated at a nominal output of 6 watts each at 118 MHz. With the feedback networks shown, the output is approximately 2 watts on 80, 40 and 20 meters. Somewhat less output is available on 15 meters, owing to the lower output from the HW-8 on that band.

A pair of 10-ohm resistors and two miniature ferrite beads are connected from the transistor bases to ground. These components are used to discourage low-frequency oscillations. The 4-dB attenuator at the amplifier input reduces the HW-8 drive to a safe level. It is suggested that a switch be added to remove the pad during operation on 21 MHz. This will provide an amplifier output of roughly 8 watts on that band. The power output is on the order of 4 watts with the pad in the line.

Two 220-pF silver-mica capacitors are used in the collector circuit to prevent vhf self-oscillations and to lower the harmonic energy in the vhf range. The reactance of the capacitors is high enough in the hf bands to have minor effect on the amplifier power.

T1 of Fig. 1 is a broadband transformer with a 3:1 turns ratio. The transformer used in this design is homemade and is of the *conventional* variety (not a transmission-line transformer). Detailed information of the construction of this transformer is presented in the *ARRL Electronics Data Book* and in *Solid State Design for the Radio Amateur*. T1 consists of two bows of four Amidon FT-50-43 ferrite toroid cores (μ = 850) through which thin-wall brass tubing is

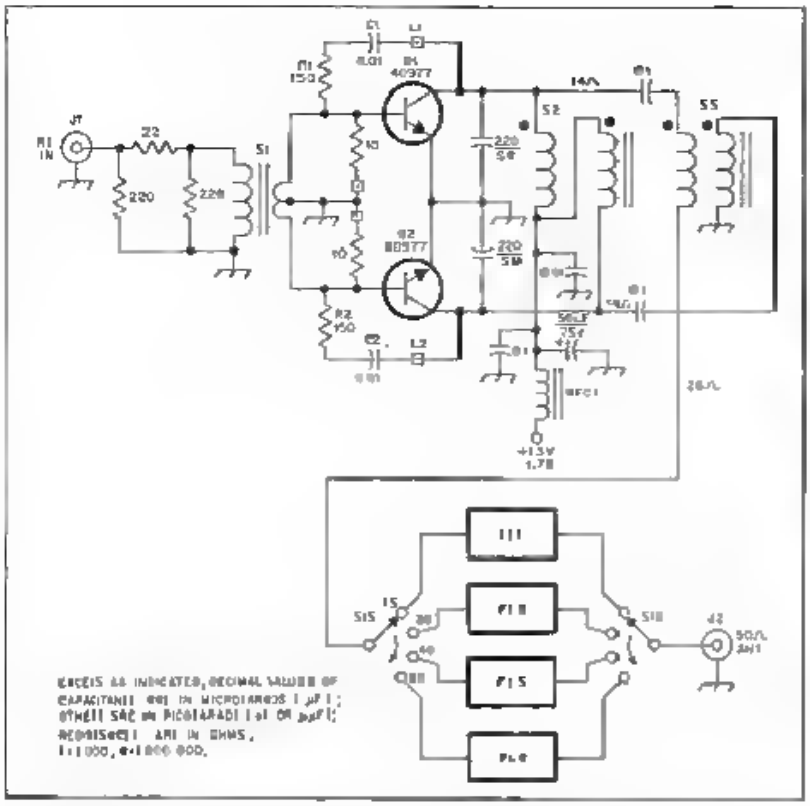


Fig 1 — Schematic diagram of the solid-state four-band amplifier. Resistors are 1/2-watt composition type. Capacitors are disk ceramic except the one with polarity marked, which is electrolytic or tantalum. S M is silver mica. Components with numbers which do not appear in the parts list are so identified for text discussion only.

J1, J2 — Panel mount coaxial connector or phono jack.

L1, L2 — Miniature ferrite bead (680 mc) over lead of 0.01-µF capacitor. Same type of beads used on pigtails of 10-ohm base resistors.

Q1, Q2 — RCA transistor (see text).

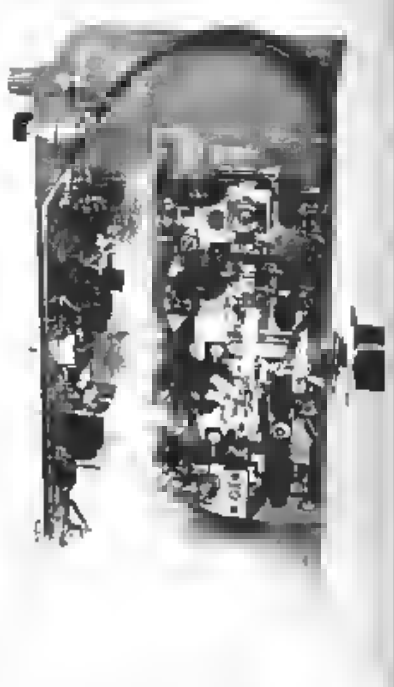
S1 — Two-pole four-position ceramic or phenolic wafer switch (see text).

T1 — Broadband 3:1 transformer (see text).

T2 — Bifilar-wound broadband phase-reversing transformer with 8 bifilar turns of no. 22 enam. wire on two stacked FT-50-43 cores (50 mc). Wires have 8 twists per inch.

T3 — Broadband current transformer with 8 bifilar turns of no. 22 enam. wire, 8 twists per inch, on stacked FT-50-43 toroid cores.

RFC1 — Toroidal rf choke, 7 turns no. 22 enam. wire on FT-50-43 toroid core.



Inside the amplifier with the filter board at right, amplifier board at left. The two power transistors are mounted against the rear panel which serves as a heat sink.

passed and made common at one end (U-shaped single turn, in effect). Then, three turns of insulated wire are passed through the tubing to form the transformer primary. Pc-board headers are used at each end of the assembly to secure the tubing and provide copper tabs for connection to the main circuit board. The advantage in using this type of transformer is that a more precise secondary center tap can be established than is possible with a simple toroidal broadband transformer. Symmetry of the secondary helps to assure equal driving power to each transistor. Those wishing to experiment with a toroidal type of transformer at T1 can wind 9 turns of no. 28 enameled wire on an FT-50-43 core. A three-turn center-tapped secondary winding can be wound over the nine-turn primary. The tap must be in the *exact* center. The leads to the transistor bases should be laid out symmetrically and have equal length.

T2 is a phase-reversal transformer that places the collectors of Q1 and Q2 in push-pull. The collector voltage is supplied through T2. A combiner transformer, T3, provides a 28-ohm output impedance from the two 14-ohm collectors. Half-wave harmonic filters (FL1-FL4, inclusive) are band switched at the amplifier output by means of S1. They are designed for a loaded Q of 1. The input impedance is 28 ohms and the output impedance is 50 ohms. Since these are low-pass filters, the cutoff frequency is set slightly above each amateur band to minimize insertion loss. Amidon powdered-iron toroid cores are used for the filter inductors.

Construction Notes

Double-sided pc board is used for the amplifier module, but single-sided board is specified for the filter assembly to reduce stray capacitance. The latter could degrade the filters by virtue of detuning effects.

Double-sided board material is used for the amplifier circuit to help eliminate ground loops and subsequently improve stability. The ground returns for the input components which relate to the transistor bases are connected to floating pads on the etched side of the board. Small pieces of wire connect those pads to the ground-plane surface on the opposite side of the board. Conversely, the collector components have their ground connections on the *etched* side of the board. This procedure helps to insure stability by breaking up rf current loops on the ground elements of the pc board. Fig. 2 shows the pc board pattern for the amplifier. Fig. 3 contains the layout for the filter board.

The photograph shows the collector-base feedback networks being bridged in mid-air over the tops of Q1 and Q2. The board pattern provides copper pads for these components. Also, the prototype version shown photographically has the transistor strip leads bent down slightly to mate with the related pc pads. This mounting technique is not recommended. Two unfavorable conditions can result from this method: Excessive emitter-lead length introduces unwanted inductance, which in turn causes degenerative feedback. Degeneration lowers the amplifier gain and may encourage instability. Fur-

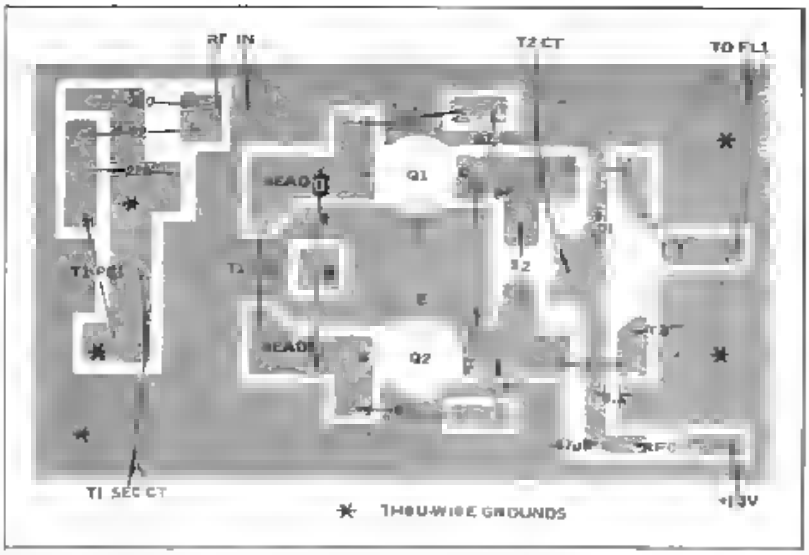


Fig 2 — Parts placement guide for the amplifier board. Parts are mounted on the etched side of the double-sided pc board, the shaded area in this view represents the copper pattern. The other side of the board is unetched. Decimal-value numbers alone represent capacitance in microfarads. Whole-number values without units represent resistance in ohms. Note that ferrite beads are slipped over one end of each of the two 10-ohm resistors.

thermore, when the leads are bent up or down to mate with the pc board it is possible for undue stress to be exerted on the transistor body during heat cycling. This can cause physical damage to the transistors. The correct mounting procedure calls for the strip leads to come out from the transistor body at 90 degrees. They lie flat on the pc board pads to which they are soldered.[2]

The amplifier board is mounted against the rear wall of the U-shaped homemade chassis. The case serves as a heat sink. Heat transfer is enhanced by the addition of transistor silicone grease. It is applied to the mating surfaces of the transistors and cabinet. The stud nuts should be tightened only slightly beyond a finger-tight tension level. This will prevent damage to the transistors. Through-wires are added at several points on the amplifier board to join the ground foils on both sides of the board. Each through wire is soldered to the pc board at both ends.

The Filter Module

Table I contains L, C and frequency data for the four filters. For the most part, standard-value silver-mica capacitors are not specified. This requires combining standard values in order to arrive at values which are close to those specified. Mica compression trimmers can be used at the center of each filter (see photograph) if desired. The author's model has the trimmers for final tweaking to obtain maximum output power and waveform purity.

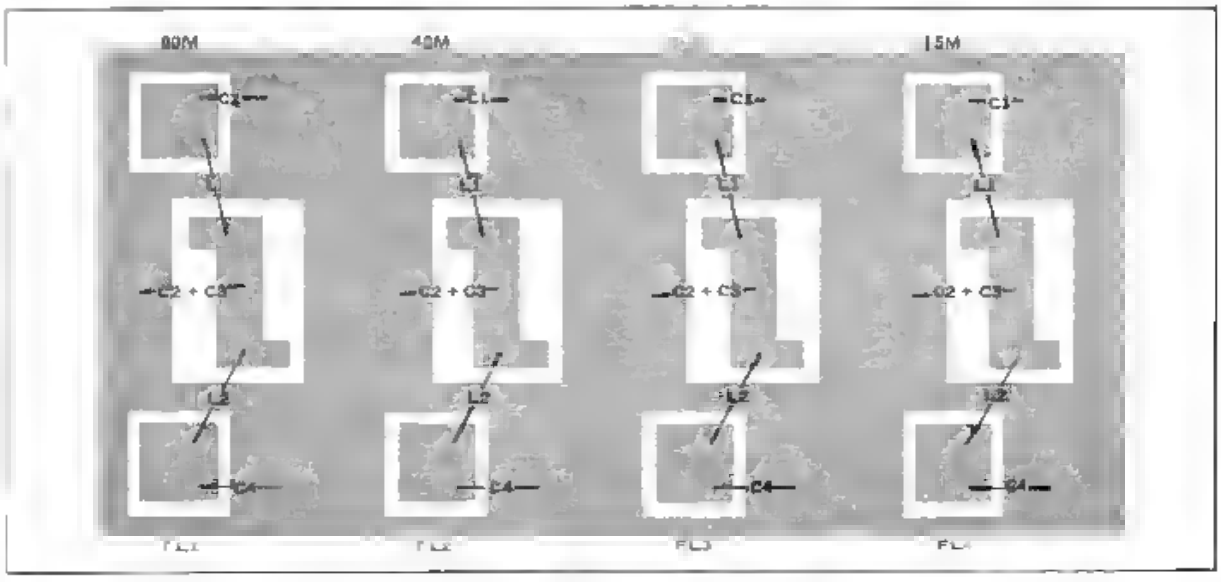
RG-174/U miniature coaxial cable is used for the rf leads. It is important to ground the shield braids at both ends of the cables which connect to the amplifier output, antenna jack (J2) and the two poles of S1. The remainder of the coaxial cables need to have the shields grounded only at the filter-board end. In the model shown, heat-shrink tubing is used at the ungrounded ends of the connecting cables. S1 should be a two-wafer type with at least one inch (25.4 mm) of distance between the wafers. This will ensure proper isolation between the filter inputs and outputs. For ideal conditions, a metal shield could even be installed between the wafer sections and bolted to chassis ground.

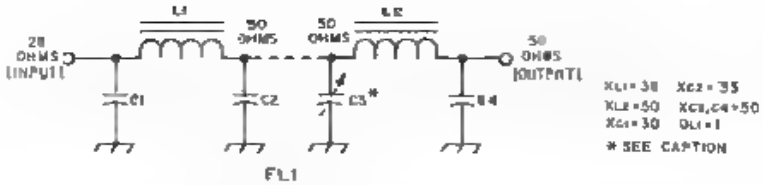
Operation

The power supply which Heath provides for the HW-8 will not be suitable for this amplifier. A regulated power supply of 2 amperes or greater is required.

The 40977 transistors are "SWR

Fig 3 — Parts placement guide for the single-sided filter board showing details for one filter. Parts are mounted on the foil side of the board; the shaded area in this view represents copper.





FL1

Table 1

Band	f_{CO} (MHz)	L1 (μH)	L2 (μH)	C1 (pF)	C2 (pF)	C3, C4 (pF)	Toroid Core	Turns	Wire Size
15	23	0.26	0.35	230	210	130	T68-6	L1-7 TS L2-10 TS	no. 22
20	18	0.38	0.5	330	300	200	T68-6	L1-10 TS L2-10 TS	no. 22
40	8	0.76	1.0	663	600	400	T68-2	L1-11 TS L2-15 TS	no. 22
80	5	1.2	1.6	1000	955	636	T68-2	L1-14 TS L2-17 TS	no. 22

Coil and capacitor information for the half-wave harmonic filters shown in Fig. 1. All capacitors are silver-mica units. Parallel or series combinations can be used as needed to provide the approximate values listed above. An accuracy of ±10 percent is recommended. *A mica compression trimmer can be used at C3 to provide final adjustment of the filters.

protected," to use the RCA vernacular. This means that anything from a dead short to a full open circuit can be tolerated at the amplifier output for short periods of time without causing device damage. A *maximum* mismatch period of 30 seconds is recommended.

The harmonic filters are designed for a 50-ohm termination. Therefore the antenna should not present an SWR of greater than 1.5:1, or filter performance will be impaired. Also, there will be a loss in output power when the SWR is high. A Transmatch and SWR indicator are recommended for use with any solid-state amplifier, including this one, particularly when the antenna does not present a 50-ohm load.

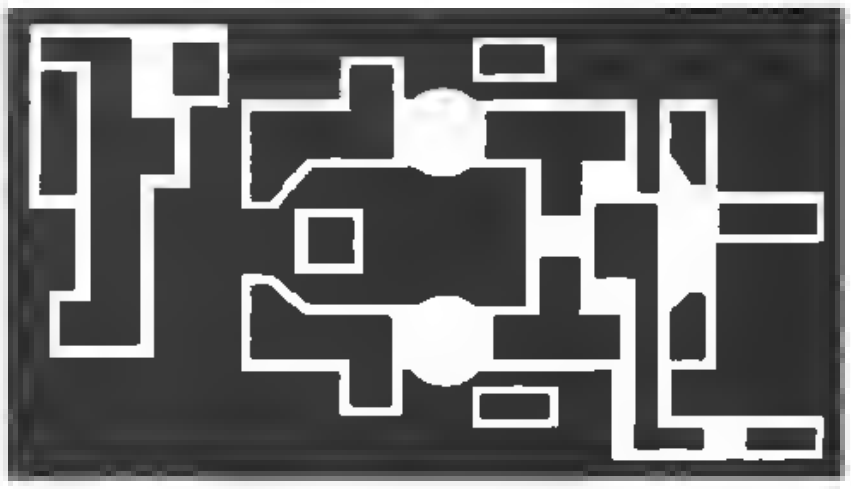
This amplifier will reach its saturated output-power level at slightly under 1 watt of drive at the bases of Q1 and Q2. Observe the increase in forward power to the antenna, then add no further drive once the point is reached where power output from the amplifier levels off.

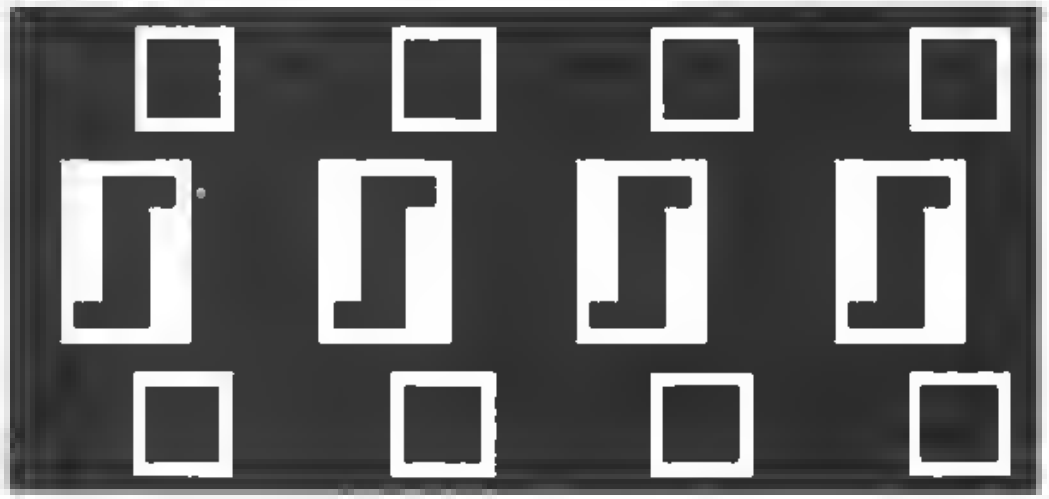
Now that you've "put the books to your HW-8," have fun and go after that DX you were reluctant to call with only 2 watts!

Footnotes

DeMaw, "Slippers for the HW-7," *QST*, Dec. 1971, p. 45.

Artigo and Johnson, "Fundamentals of Solid-State Power-Amplifier Design," *QST*, Sept. and Nov. 1972, and *QST*, April 1973 (in three parts).





Circuit-board etching patterns. The front sides of the boards are shown here at actual size with black representing unetched copper. The upper pattern is for the amplifier section (Fig 2). It is copper clad on both sides with unetched copper on the "back" side that forms a ground plane. The lower pattern is for the filter board (see Fig 3).

From May 1984 QST, p 44:

30-Meter Conversion For The HW-8

☐ The Heath HW-8 QRP transceiver can be modified easily to operate on 30 meters if you are willing to sacrifice one of the existing bands. I chose to give up the 80-meter band, since I have found it to be the most demanding one, in terms of antenna size, for QRP operation. Thirty meters seems to be an excellent band for QRP operation, and it offers the side benefit of WWV reception, which I use to calibrate my VFO dial.

Complete details of the modification are summarized in Table 1. The only expensive component is the crystal, which costs around $10. The other components can be found in your junk box or purchased from a variety of *QST* advertisers. Five of the original capacitors are reused in other locations.

Remove the control knobs and front panel; then, disconnect the loading capacitor from the front of the chassis. This will make it easier to get at the components to be changed in the crowded area around SW1 (the 80-meter band switch). Remove the indicated components using a vacuum desoldering tool, solder wick or a piece of flattened braid from coaxial cable.

After the new components have been installed, the rig can be aligned according to the instructions in the HW-8 assembly manual. The only problem I encountered was that I had lost the small tuning tool used to adjust L17 in the heterodyne oscillator. I found that the larger tool of even an Allen wrench can be used. Carefully insert the tool through the top slug and tune the bottom slug for maximum output on 30 meters. Then, back the tool out and readjust the top slug (L18) for maximum output on 40 meters.

The transmitter dc power input should be about 3 W. The VFO will cover 10.0 to 10.25 MHz. Dial accuracy seems to be a problem with the HW-8, so it may be difficult to determine the band edges without a frequency counter. This is where WWV can assist you. Just be sure to stay within the legal segments (10.100-10.109 and 10.115-10.150 MHz). If in doubt, don't transmit.

—Wayne Burdick, N6KR 7904 Caminito Dia #2, San Diego, CA 92122

Table 1
HW-8 30-Meter Modifications[1]

Part No.	*New Value*	*Description*
Yc	18.095 MHz	Fundamental type, 15-pF load, HC-6/U holder. International Crystal Mfg. Co., P.O. Box 26330, Oklahoma City, OK 73126. Part no. 4J411Z.
L1	1.8 µH	Secondary — 25 turns no. 24 enameled wire on T37-6 core (Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607). Primary — 2 turns no. 24 wire over G2 end of secondary (use original coil form).
L5	1.8 µH	25 turns no. 24 wire on a T37-6 core.
L13	4.0 µH	Remove 16 turns from original L13.
L22	2.7 µH	23 turns no. 22 wire on a T50-2 core.
L26, L27	3.2 µH	25 turns no. 22 wire on a T50-2 core.
C1	100 pF	Silver mica, 5% tolerance (use original C116).
C15, C86	100 pF	Silver mica, 5% tolerance.
C64	68 pF	Silver mica, 5% tolerance (use original C91).
C77	220 pF	Silver mica, 5% tolerance (use original C64).
C78	150 pF	Silver mica, 5% tolerance (use original C90).
C84	47 pF	Silver mica, 5% tolerance (use original C15).
C91	300 pF	Silver mica, 5% tolerance.
C118	39 pF	Silver mica, 5% tolerance.
C301A	—	Disconnect from L1.
R50	—	Remove.
R56	1 kΩ	½ W, 10% tolerance.

[1]Refer to HW-8 schematic diagram for part locations.

From April 1988 QST, p 26:

Improving the HW-9 Transceiver

If you own an HW-9 or other QRP transceiver, you'll find these ideas will add to your operating enjoyment. So, heat up that soldering iron!

By Chuck Hutchinson, K8CH and Zack Lau, KH6CP
ARRL Technical Department

This article is divided into two parts. In the first part Chuck, K8CH, describes the portable QRP station that he uses for Field Day and vacation operating. The second part describes circuit modifications by Zack, KH6CP. Although the ideas presented concentrate on using and improving the Heath HW-9, they can be adapted to many QRP rigs.

Chuck's QRP Package

I enjoy chasing DX with QRP—most of the time. But Field Day and vacation are two times when QRP operation is particularly appropriate and rewarding. I'm not averse to running 100 watts (or even the legal limit when conditions warrant), but my entire QRP station with transceiver, power supply, antenna, keyer and other accessories is about the same size and weight as my 100-W, full-feature transceiver. That means it's a lot easier for me to take the QRP station to the Field Day site. As for vacation, only the QRP rig will fit into the car along with the rest of the family luggage.

I use two 9-Ah gelled-electrolyte, lead-acid batteries as a portable power supply.[1] These are not lightweights, but they're good for many hours of operation. Exact time before recharging is required depends on duty cycle. In other words, transmitting "eats" the batteries more rapidly than receiving.

While one battery is powering the transceiver, the other can be recharging. My favorite method of recharging the batteries is to use a solar panel—mine is rated at 18 V and 500 mA.[2] It feels good to put those free photons to work—and solar energy is good for bonus points on Field Day! An ac-operated charger was described in June 1987 *QST*.[3] That charger ensures optimum charging of batteries. For best battery life, don't run the batteries flat before recharging. *The ARRL Handbook* explains proper care of lead-acid batteries (Chapter 6 in recent editions).

[1]Notes appear at end of article

Portable Antennas

For portable operation, I like to use a dipole suspended by rough, lightweight nylon cord. The dipole in my portable station uses plastic insulators (see Fig 1). The center insulator has an extra hole so that a nylon line can be used to support the

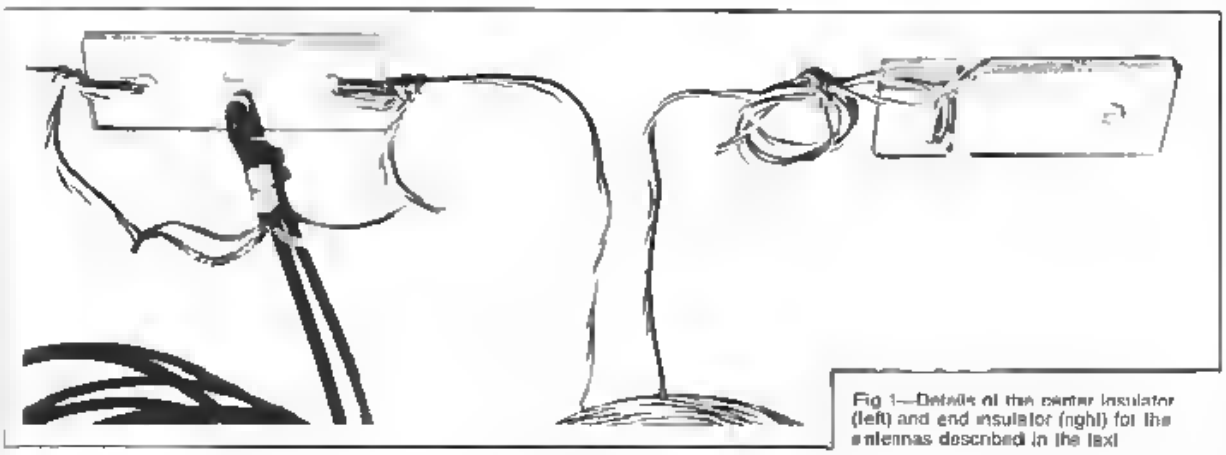


Fig 1—Details of the center insulator (left) and end insulator (right) for the antennas described in the text.

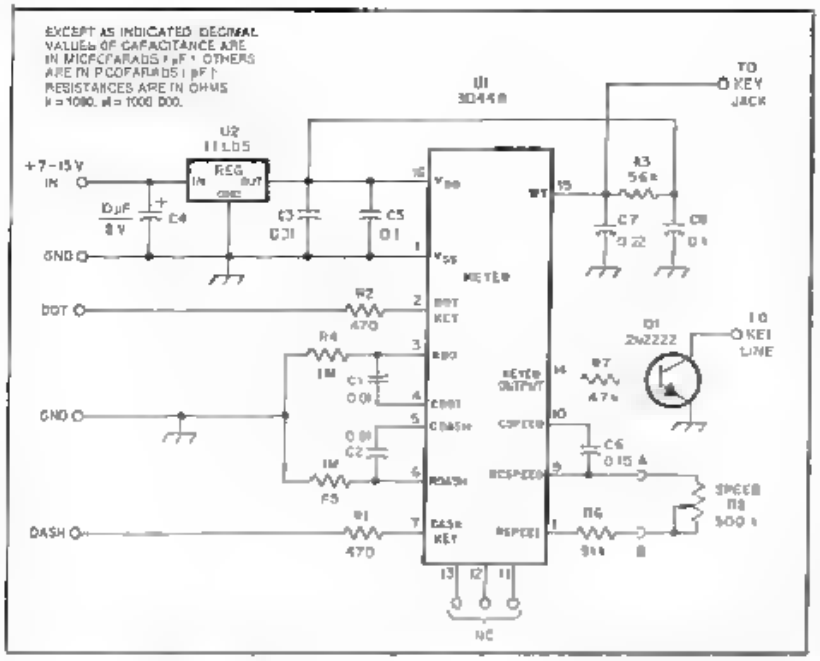


Fig 2—Schematic diagram of a keyer based on the Curtis 8044 IC. Capacitors are disc ceramic, except for C4, which is electrolytic. C6 and C7 are NP0 types, although any temperature-stable capacitor of the proper value should work fine.

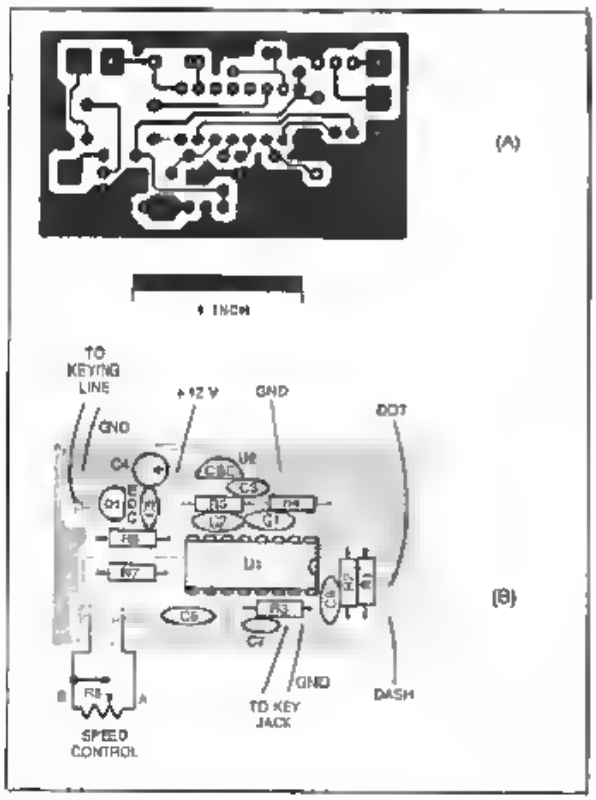


Fig 3—Circuit-board etching pattern (A) and parts-placement guide (B) for the keyer. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern.

center. The end insulators are made so that element lengths can be adjusted easily for changing bands.

For 80-meter operation, I use an end-fed quarter wavelength of wire terminated with a banana plug. The plug fits neatly into the RF connector on the HW-9's rear panel. The far end of the wire is supported by a plastic insulator like those used in the dipole. Because this antenna operates against ground, I carry a couple of clip leads to make a connection to the best ground I can locate. (For instance, I've had good luck grounding to the heating pipes in a motel. The secret is to use what you have available.)

Accessories

At first, I used my son Scott's (N1DSF) Heath µMatic Memory keyer with the HW-9. (I mounted a phono connector to the HW-9's rear panel to provide switched 12 V dc power for the keyer.) Later, I decided to build into the transceiver a keyer based on the Curtis 8044 CMOS IC. The circuit is based on the 8044 spec sheet, and the schematic is shown in Fig 2. Tom Miller, NK1P, prepared the schematic and the PC board shown in Fig 3. I mounted the completed board upside down using a bolt and nut that holds the HW-9's BFO shield in place. I moved the wire from the key jack to the circuit board, and ran a new wire from the circuit board to the key jack. The SPEED control, R8, is added to the front panel, and a jack for the paddle is added to the rear panel.

For portable operation, I wanted to package the station for easy transport. An aluminum briefcase proved to be just what I was looking for. Packing foam, cut with a hacksaw blade, cushions the HW-9. The rest of the station, except the solar panel, goes into the case with the HW-9: the two gel batteries, dipole with feed line, 80-m end-fed antenna, nylon cord, clip leads, keyer paddle, lightweight headphones and an ARRL Minilog.

Conclusion

My portable QRP station is not made for backpacking. It does, however, fill my need for something that goes easily to Field Day or on vacation. The entire station, except for the solar panel, fits into a briefcase. What could be more convenient?—*Chuck, K8CH*

Zack's Circuit Improvements

Although this portion of the article concentrates on improving the Heathkit HW-9 QRP transceiver, these modifications may be of general interest to home-brewers, as they can be adapted to many QRP rigs. These modifications include adding an SWR meter that requires no balancing adjustments, removing audio thumps and clicks, and improving the signal-to-noise ratio of the HW-9's NARROW audio filter.

The new HW-9 SWR meter is a version of the directional coupler used in the Tandem match.[4] The main advantage to the coupler shown in Fig 4 is that no adjustments are required. Anyone who has fiddled with trimmer capacitors trying to get a good null will appreciate this feature. Faraday shielding is not used in this application, as coupler directivity is adequate for the uncompensated diode detectors.

The switching circuit, shown in Fig 5, allows the existing HW-9 meter to be used as an SWR meter on transmit and as an S meter in its normal function) on receive. When the voltage at the input of this circuit (Q403 collector) is zero, Q1 turns on and Q2

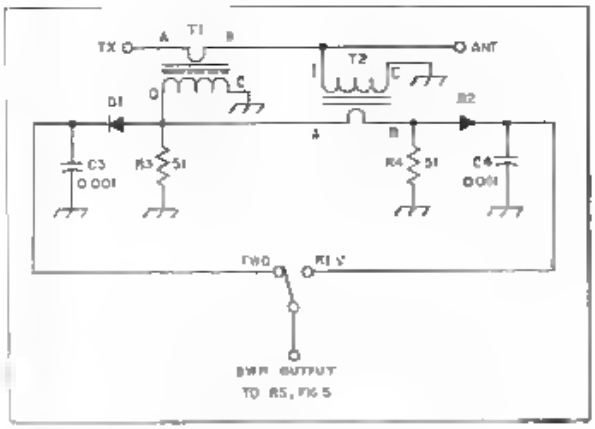


Fig 4—Schematic diagram of the SWR meter
D1, D2—Schottky diode HP 2800-2535 (1N34A can be substituted)
T1, T2—Broadband transformer. Primary has 14 turns of no. 28 enam wire on an Amidon FT-23-43 toroid core. Secondary is 1 turn of no. 22 hookup wire

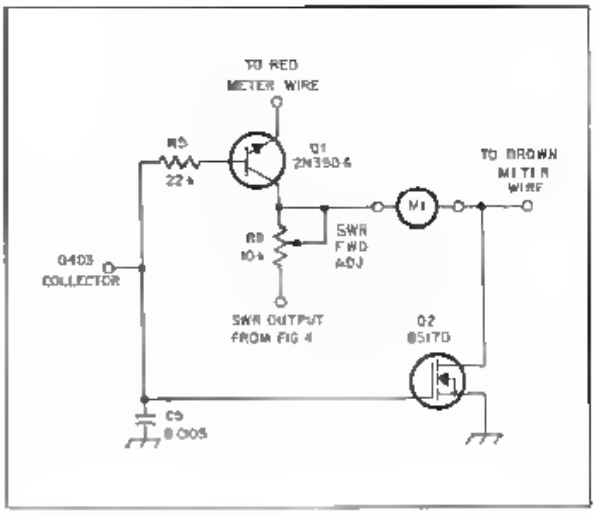


Fig 5—Schematic diagram of the meter switch circuit. Q403 is a Heathkit part designator.

turns off. This allows the meter to function normally. When the input voltage is raised to 12 volts, as is the case during transmit, Q1 is turned off and Q2 is turned on. Q1 now prevents current from the S-meter circuit from affecting the SWR measuring circuit. When Q2 is turned on, it effectively shorts out the S-meter calibration voltage, as it is not wanted while using the meter to measure SWR.

The audio thump suppressor is used to reduce the audio thumps that result when the HW-9 switches from transmit to receive. The audio line in the original HW-9 sounds like it's being shorted out when the rig switches between transmit and receive because a transistor, Q303, is used to do exactly that! A 12-dB reduction in audio thump can be obtained by using a JFET switch to break the audio line while transmitting. See the schematic in Fig 6. When the gate of the JFET Q3 follows the source, the JFET acts as a resistor with a value of roughly 100 to 300 ohms. When the gate is grounded, the JFET effectively breaks the audio line. A dc bias of roughly $V_{cc}/2$ is needed at the source of the JFET for the circuit to work. This is supplied by the output of U304. C2 is used to reduce the high-frequency response of the switch to help remove the high-frequency audio clicks. R2 is optional. A properly selected value for R2 will provide a degree of audio limiting and further thump reduction beyond the measured 12 dB. It is possible to eliminate the thump entirely by adding additional low-pass or band-pass filtering after the JFET switch. The remaining thump exists only in the wide filter position, as the narrow filter removes it.

The final modification increases the dynamic range of the HW-9 by a few decibels. If the capacitor values in an active filter circuit are too small, a substantial increase in noise results. The new values shown in Table 1 are chosen for a 250-Hz Bessel response centered at 700 Hz. A Bessel response is chosen to eliminate ringing. Measurements in the ARRL lab indicate that the filter shape tends to change at very low signal levels if the capacitor values are too small. In some cases, the band-pass response actually becomes a notch response, although the notch is usually above the desired passband.

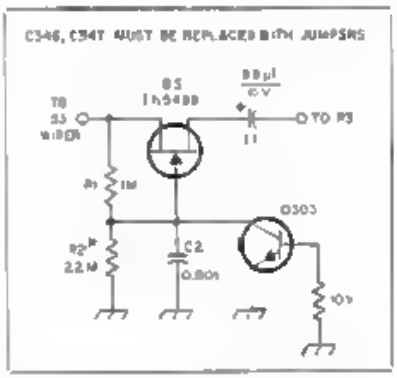


Fig 6—Schematic diagram of the audio thump suppressor. Q303 is part of the transceiver—see text.

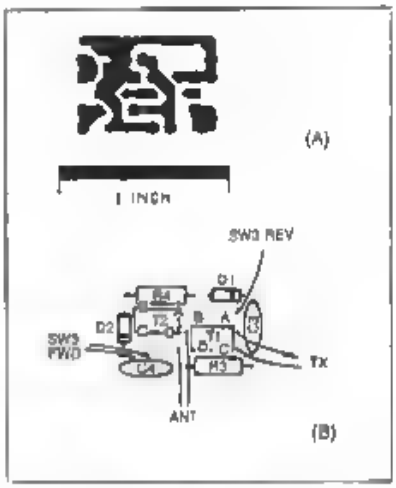


Fig 7—Circuit-board etching pattern (A) and parts-placement guide (B) for the SWR meter. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil. Parts are placed on the nonfoil side of the board. The shaded area represents an X-ray view of the copper pattern.

Construction

The modifications to the HW-9 involve stuffing two PC boards and changing parts on the TR circuit board. The board shown in Fig 7 contains the directional coupler, and the board shown in Fig 8 contains the audio-thump suppressing circuit and the meter-switching circuit. This allows the coupler to be mounted in the back of the rig next to the antenna jack, while keeping the thump removal circuitry next to the audio section.

It is essential that C346 and C347 be replaced with wire jumpers for the thump removal circuit to work, as they would block the needed dc bias voltage. One of these capacitors can be used as C1 on the modification board, but take care to get the polarity right.

Thanks to Heath's excellent design, it is not necessary to unsolder all the wires to get to the solder side of the TR board. First,

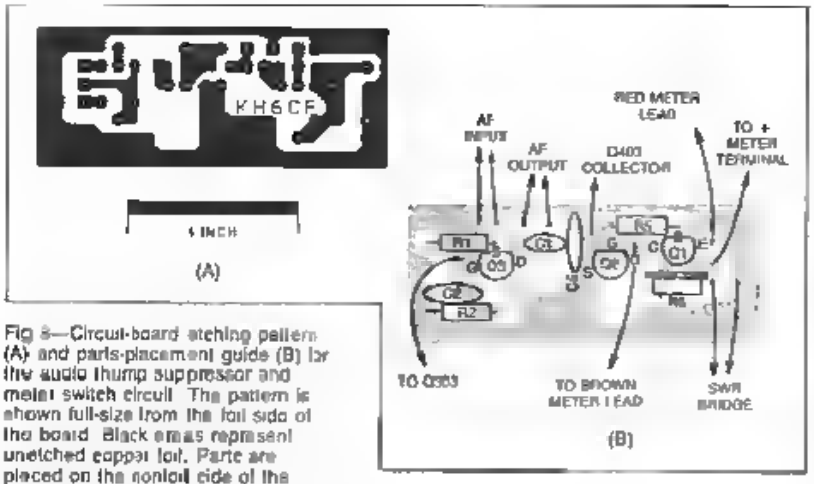


Fig 8—Circuit-board etching pattern (A) and parts-placement guide (B) for the audio thump suppressor and metal switch circuit. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil. Parts are placed on the nonfoil side of the board, the shaded area represents an X-ray view of the copper pattern.

set the BAND switch to 20 meters and remove the BAND-switch shaft. Then, unsolder the blue keying wire and remove the six screws holding the back panel to the chassis. After removing the five nuts securing the TR board, the circuit board can be flipped up, exposing the foil side. While you have the foil side of the TR circuit board exposed, install the NARROW audio filter components listed in Table 1. I used metal film capacitors, but polyester or polystyrene capacitors can also be used, although the latter may be physically a little large. I recommend using 5%-tolerance resistors to prevent the center frequency of the filters from being too far off.

The collector lead of Q303 has to be unsoldered and attached to a hookup wire that goes to the thump suppressor. This "flying mod" is unsightly, but I see little alternative. A 10-kΩ resistor must be soldered between the base of Q303 and ground, or the thump suppressor may not allow any audio through! The resistor prevents quiescent currents from keeping the transistor on when it isn't supposed to be.

I made the switch for choosing forward or reverse power readings by combining it with the existing audio SELECTIVITY switch. First, I bought a standard Switchcraft DPDT slide switch. The plastic slider handle is too short, so I then bent the metal tabs holding the (new and old) switches together to take them apart. I then swapped the plastic slider handles, taking care not to lose the metal slide contacts. This gave me a DPDT switch with a long slider handle. You could mount a separate switch if you like, but I prefer modifications that don't require making holes in the front panel.

I used RG-174 cable on the audio and SWR meter connections to prevent unwanted signal pickup. The rest of the connections are made with standard hookup wire.

Table 1
Component Changes

Part No	Old Value	New Value
R352	680 k	8 k
R353	33 k	6 k
R354	1.5 M	39 k
R357	680 k	5.6 k
R358	33 k	680
R359	1.5 M	18 k
C339, C341	1000 pF	0.027 µF
C344, C345	1000 pF	0.082 µF

Once you're sure everything is installed correctly, turn on the rig. Hopefully, the audio hiss in the WIDE position will be as loud as before. If not, Q3 may have been installed backwards, or you may have forgotten to replace C347 with a jumper. The hiss should be less in the NARROW position, because the modification is supposed to reduce noise. With the rig hooked up to a dummy load, you should be able to notice much less of an audio thump when using WIDE audio selectivity, and no thump when using NARROW. If a nasty thump is heard, Q303 is not hooked up properly. If you hear just a little bit of thump, you may consider adding R2 to reduce the thump by a few more decibels. Basically, you want as low a value of R2 as possible without turning the audio off all the time. Typical R2 values range from 1.5 to 2.2 M, depending heavily on the FET used.

While transmitting into a dummy load, adjust R6 for the desired meter deflection in the forward position. If the meter deflects the wrong way, a diode is hooked up backwards. A bad Q2 (power MOSFET) will either affect the S-meter calibration or make the bridge read backward with no power output. A properly operating bridge will measure little, if any, reflected power when using a dummy load. Since there is no SWR specification for the MRF-237s, you should be careful not to transmit into a load with an SWR greater than 2.

When all is working well, reassemble your HW-9 and enjoy. I'm sure you'll find your transceiver more pleasant to operate.—*Zack, KH6CP*

Notes

[1]Gelled-electrolyte 9 Ah batteries are available from American Electronics, 173 E Broadway, Greenwood IN 46142, tel 317-888-7265. Reliance Dick Smith part no. S-3321, price $34.95. American Electronics also sells a charger that operates from 120 V ac. Reliance Dick Smith part no. M-9523, price $9.95. For shipping and handling add $1.50 plus 5% of order. American Electronics has a $20 minimum order.

[2]A solar panel rated for 1 A at 9 V or 500 mA at 18 V is also available from American Electronics. Reliance Dick Smith part no. Z-4845, price, $149. See note 1.

[3]Warren Dion, N1BSH, "A New Chip For Charging Gelled-Electrolyte Batteries," *QST*, Jun 1987, pp 28-29.

[4]John Grebenkemper, KA3BLO, "The Tandem Match—An Accurate Directional Wattmeter," *QST*, Jan 1987, pp 18-26.

From October 1988 QST, p 43:

HW-9 Tips

I read the article, "Improving the HW-9 Transceiver,"[1] with great interest. I built an HW-9 about two years ago, and the first thing I added to it was a Curtis keyer chip; the second thing was a 100-kHz crystal calibration oscillator. The keyer and calibrator oscillator circuits are contained on a small perf board that's secured to the left rear corner of the rig by means of small metal angle brackets. The calibrator has proved extremely useful in light of the HW-9 reduction drive's tendency to slip.

I'd like to add a couple of suggestions concerning the addition of the keyer circuit. I like to use a straight key from time to time, so I removed the original key jack, enlarged the hole and mounted a four-pin microphone connector in its place (see Fig 2). This provides connections for both a paddle and straight key without adding another jack. I also added a small push-button switch to the rear panel and connected it to the keying line for use as a TUNE switch.

I found the HW-9's keying to be a bit on the heavy side. Although the weighting could have been altered by using a weighting control connected to the Curtis chip, I decided it was better to correct the problem at its source: This is the HW-9's keying line, which has a slow return to +12 V. I solved this by adding a 1-kΩ resistor from the transmitter keying line to +12 V. Observation of the rig's output on a scope shows almost perfect weighting. I recommend this simple modification to anyone using an external keyer as well.

I'm already planning my next project: Add the SWR meter, thump suppressor and filter modifications described in the April article. With these additions, this great little rig will be even more of a joy to operate! Now—if I could just find a way to reduce the warm-up drift of the VFO. —*Larry V. East, W1HUE/7, POB 51445, Idaho Falls, ID 83405-1445*

[1]Deleted

[1]C. Hutchinson and Z. Lau, "Improving the HW-9 Transceiver," *QST*, Apr 1988, pp 26-29.

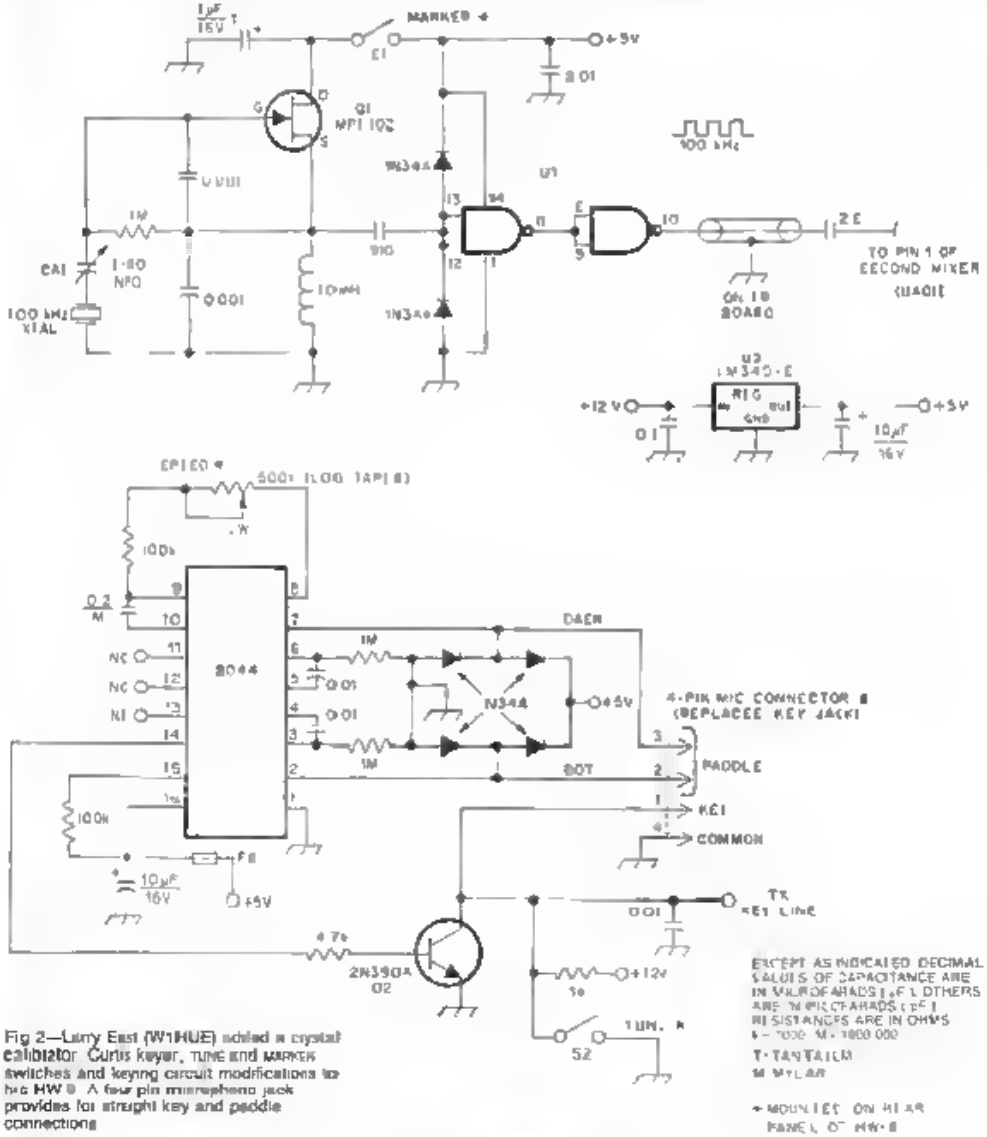
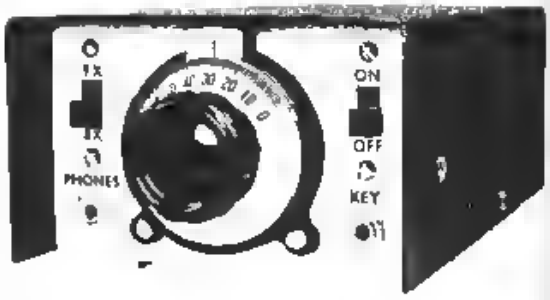


Fig 2—Larry East (W1HUE) added a crystal calibrator, Curtis keyer, TUNE and MARKER switches and keying circuit modifications to his HW-9. A four-pin microphone jack provides for straight key and paddle connections.

From June 1975 QST, p 35:

The Mavti-40

Part I



BY D. K. SIEMER,* K0IYD

We have had many requests for a simple transceiver that is within the building capabilities of most beginners. The MAVTI-40 described here is the ideal answer to these requests. As the author points out, this is not a one of a kind unit, as several have been built by his students, and they all work

THE 7-MHz TRANSCEIVER described in this article is the result of a desire to have a small, portable station for personal use. Also, since many of the friends and members of the Mankato Area Vocational-Technical Institute are interested in ham radio and are mainly short of funds, it seemed like a good idea to make it inexpensive enough to be available to the most enthusiastic beginners. They can build the receiver section first for code practice, then they can build the transmitter later when they get their tickets.

This project makes use of new components rather than surplus ones to make parts procurement easy and to avoid the pitfalls and disappointments often associated with the latter. The components, though new, are not expensive, and the whole unit can be built for $40 to $50 – less key, cabinet, and equipment included. Several of these units are now in operation and have produced many satisfied-operator reports.

Building economy was kept in mind while the transceiver was built with a minimum number of components consistent with good design and satisfactory operation. None of the units built have exhibited unusual problems in construction or operation, making the MAVTI a good candidate for a first homemade project.

The VFO

The VFO for a variation of one used in a number of projects here, Q1, Fig. 1, performs as a Colpitts oscillator and Q2 as a source-follower buffer. To keep parts to a minimum and still have good mechanical stability with high output voltage, a toroid core was used with L1 instead of the usual single-layer solenoid one. C5 is a temperature-compensating capacitor to reduce oscillator drift.

When tested in an environmental chamber, the output frequency shifted less than 2 kHz with a temperature variation between 50° F and 100° F. Below 50° F the frequency shifted quite rapidly however, typically 1000 Hz/° F which would be of concern if low-temperature operation is anticipated. Frequency shift between transmit and receive is less than 200 Hz and warm-up drift is less than 150 Hz in the two-minute period immediately after turn-on. After the two-minute period the oscillator drift is so slight it is unnoticeable.

The tuned-circuit component values were chosen so that the tuning capacitor, C4, and it just covers the 150-kHz cw portion of the 40-meter band with a little left to spare. C3 is a trimmer capacitor to adjust the oscillator frequency to 7000 kHz with C4 fully meshed. The 5 to 25-pF trimmer given in the parts list would be more satisfactory than the 3 to 12-pF value shown in Fig. 5. C3 is mounted directly on the solder lugs of C4.

The MFC4060A voltage-regulator chip offers superior performance when compared to a Zener diode. Line regulation is typically 0.1% per volt. Since oscillator supply voltage regulation was an important consideration when designing the unit.

The VFO circuit board was laid out so that a Mallory 2-inch vernier dial could be bolted directly to it. The two hex nuts are on the 4-40 × 1-inch mounting screws that hold the dial to the front panel. Two 5/8-inch long standoffs hold the board away from the front panel the proper distance for connecting the capacitor shaft to the vernier dial. A small 1/4-inch long bushing 1/4-inch OD and 3/16-inch ID is used between the capacitor shaft and the vernier drive shaft. The bushing was made by drilling a 3/13-inch hole through a 1/8-inch brass rod and cutting it to length. It was then slitted along one side to allow the bushing to compress against the shaft of C4 when the dial drive setscrew was tightened against it. To insure mechanical stability, L1 was glued to the pc board by means of silicone rubber adhesive.

The Receiver

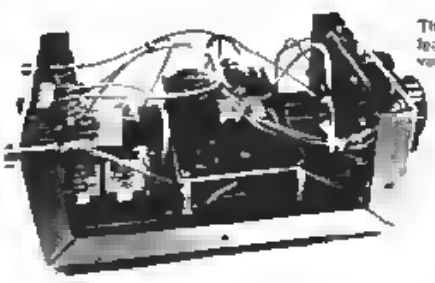
The receiver section (Fig. 1) makes use of a MOSFET, Q3, in a straight forward direct conversion scheme as described in numerous technical articles as well as the ARRL *Handbook*. The unusual component is the toroidal transformer consisting of L5, L6, L7 and C16.

Windings L5 and L6 make up a 2.25 : 1 step-down impedance matching transformer between the drain of Q3 and the base circuit of the audio amplifier, Q4. L7 and its associated capacitor C16 in 2.2 µF. 3-volt disk ceramic in the area, provides a transformer resonance in a useful frequency of approximately 800 Hz with a bandwidth of 100 Hz. This transformer is wound on a cup-core assembly consisting of two cup-shaped pieces of ferrite material that surround a nylon bobbin.

While the cup-core transformer is not very common in amateur work, it is widely applied in industry where high Q, compact, self-shielding inductors are required. The parts needed for this assembly may be obtained from Linn Ferrite Laboratories, whose address is given in the parts list (see Fig. 1). Be sure to order two of the cup cores and one bobbin as they are not sold as an assembly.

The bandwidth of the transformer can be varied by changing the reactance of L7 and shunting it with C16. In the first unit built, L7 was 70 turns of No. 30 AWG and C16 was 0.68 µF. These values provided a bandwidth of about 400 Hz centered on 800 Hz. C16 should be a low-loss type with ceramic, mylar or polystyrene dielectric.

For individuals interested in experimenting with the cup-core transformer, the 3B7-1000-3C19F material used here has an unmodified inductance value of A_L of 7580 mH/1000 turns. Different



This shows the inside of the transceiver. Shielded leads are used for all interconnections and to the various terminals on the rear of the enclosure.

*c/o V-Tek Inc., PO Box 3104, Mankato, MN 56002

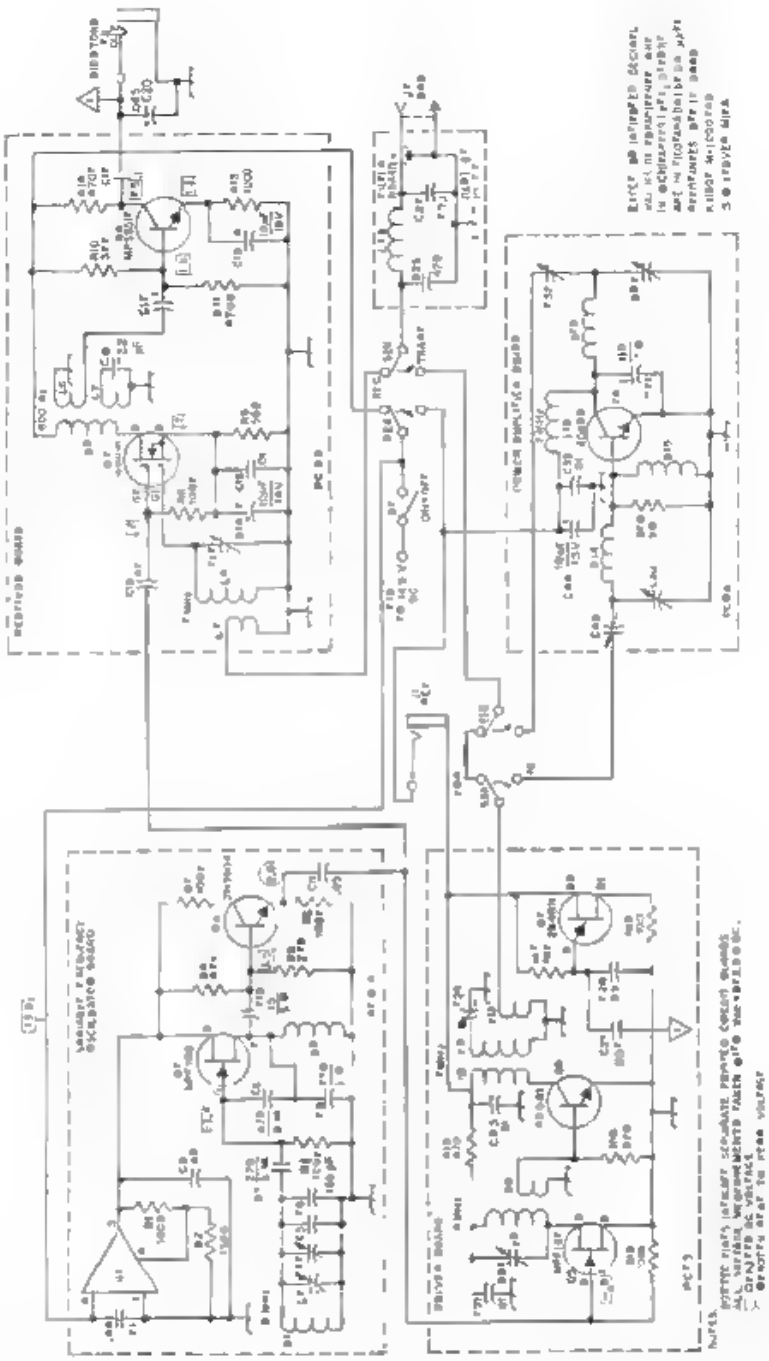


Fig. 1 — Circuit diagram of the MAVTI 40-meter portable. Resistors are in ohms, all resistors are 1/2 watt.

C3 — 5- to 25-pF ceramic.
C4 — 2.7- to 15.6-pF variable (E. F. Johnson 160-110-5 or equiv.).
C5 — 47 pF, N750 ceramic.
C13, C22 — 10- to 180-pF mica compression type (ARCO 463).
C24, C29, C30, C32, C33 — 75- to 480-pF mica compression type (ARCO 466).
J1, J2 — Open-circuit jack.
J3 — Phono jack.
L1 — 19 turns No. 28 enam. wound on Amidon T37-2 toroid core.
L2 — 56 µH molded rf choke.
L3 — 4 turns No. 24 enam. wound over L4.
L4 — 34 turns No. 26 enam. wound on T50-2 toroid core.
L5 — 305 turns No. 32 enam. wound on cup core, see note below.
L6 — 162 turns No. 32 enam. wound over L5.
L7 — 50 turns No. 32 enam. wound over L5 and L6.
L8 — 34 turns No. 26 enam. wound on T50-2 toroid core.
L9 — 3 turns No. 24 enam. wound over L8.
L10 — 7 turns No. 24 enam. wound over L11.
L11 — 22 turns No. 22 enam. wound on T50-2 toroid core.
L12 — 4 turns No. 26 enam. wound over L11.
L13 — 19 turns No. 20 enam. wound on T50-2 toroid core.
L14, L17 — 20 turns No. 24 enam. wound on T37-2 toroid core.
L15, L16 — 65 turns No. 32 enam. wound on T37-2 toroid core.
U1 — MFC4060A.

(Note: L5 is wound on a Ferrocube 3019P38 ungapped cup core. See text for winding details. This device is available from Fine Ferrite Labora tories Inc., P.O. Box 395, Woodstock, NY 12498. The T50-2 and T37-2 toroid cores are available from Amidon Assoc., 12033 Otsego St., North Hollywood, CA 91607.)

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

values of inductance may be calculated using the following equation:

$$\frac{L_1}{(N_1)^2} = \frac{L_2}{(N_2)^2}$$

L_1 = known A_L.
N_1 = 1000 turns.
L_2 = Inductance (known or unknown).
N_2 = Number of turns (known or unknown).

Where L_1 = 7580 mH and N_1 = 1000 turns. The equation is the same as that used with the Amidon cores used in this unit.

Although it does not have the steep skirt selectivity that a more elaborate passive or active filter may have the tuned transformer approach yields excellent results for a minimum number of component and cash outlay.

The transformer assembly is held down on the pc board with a No. 4-40 x 1-inch screw and washer through the center mounting hole. Be careful not to apply excessive torque in the screw when assembling the unit because the ferrite material is extremely brittle and may crack.

The detected audio is amplified by Q4 and then applied to 2000-ohm headphones via J2. When constructing the unit, don't forget C20 on J2, it prevents rf from being transferred to the headphone end and being reradiated into the front end of the receiver. This was a source of spurious oscillations which caused considerable grief when the circuit was being developed. The problem also showed up on a completed unit from which C20 was omitted.

The receiver has a comfortable listening level with three to four microvolts input. Agc control is controlled by positioning the headset for a comfortable audio level. For strong signals they may be laid on the table and used as loudspeaker.

The receiver board is quite versatile and can be used as a product-detector/rf-preamplifier stage in a superheterodyne circuit by changing L3, L4, and C3 to research the intermediate frequency and replacing the VFO input with a BFO of the proper frequency to produce a beat note. The board requires a Vcc jumper to operate. The jumper location on the board may be observed on Fig. 3.

In Part II of this article, we'll describe the driver and amplifier stages, pin line-up procedures. Meanwhile readers interested in constructing the unit can begin accumulating the parts shown in Fig. 2 and in the parts list.

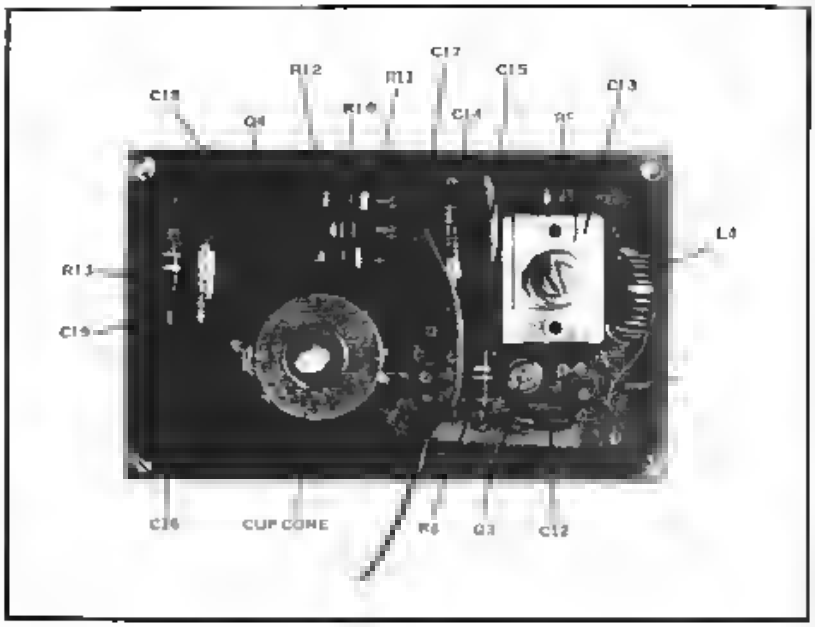


Fig. 2 — Parts placement for the receiver board.

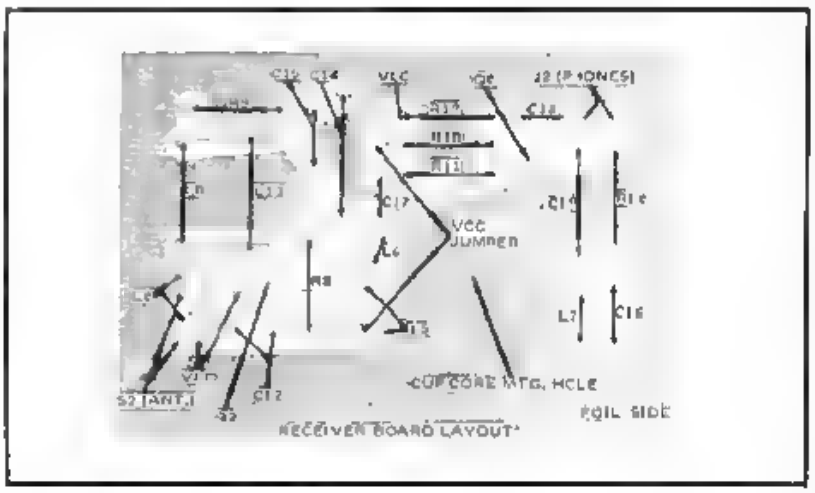


Fig. 3 — Full-size template for the receiver board.

MAVTI-40

Part II

The Driver Stage

IN PART I of the article, we described the [illegible] and VFO sections of the [illegible] transceiver. With the information provided in this section, the builder can complete the station.

The driver board [illegible] that is an adaptation of the Millywatter by W7ZOI as described on page [illegible] of the *Radio Amateur Handbook*, 1972 edition.

To minimize VFO loading, Q4, a JFET, was used as the driver transistor. Q5 drives the base of Q6, an amplifier, via L9 which is a 3-turn link wound over L8. Q6 has a typical output of 700 mW when V_{cc} is 13.6 volts. The output drops to a little over 300 mW when V_{cc} is reduced to 12 volts, making it a usable transmitter when flashlight or lantern batteries are the only available source of power. Q6 is also [illegible] efficient in this circuit.

The low-pass filter consisting of C25, C26, and L12 removes most of the harmonic energy present in the output from Q6. It is wired between the antenna terminal and the rest of the circuitry, and is effective on both high- and low-power transmitting [illegible] modes.

Q7 is a FET that forms a sidetone oscillator. Whenever the driver is keyed, Q7 turns on and its output is coupled to the headphone jack via C27. The value given for C27 provides a comfortable level of sidetone, but it may be changed to suit individual preference. C28 and R17 determine the sidetone frequency, which is typically 1 kHz. With C22 and L24 peaked at midband, 7035 kHz, the output amplitude is constant from one end of the band to the other.

The number of turns for the inductor on the driver board is somewhat critical and should be counted carefully. To make counting and hookup easier, different color wire was used for each of the windings. The wire size is not critical. However, it is best if the turns are spaced equally along the circumference of the core. Be sure that L9 is wound over the middle of L8 and not in the gap [illegible]. This can be a cause of loss of output.

Depending on the position of S3, the output from the driver board is either coupled to the antenna via the T-R switch, S2, and the low-pass filter, or to the input network opposite the power amplifier board.

The Power Amplifier

The power amplifier board was designed using the procedure given by W7ZOI in the May 1972 issue of *QST*. Excitation from the driver board is coupled to the base of Q8 via the input [illegible] consisting of C29, C30, and L14. The [illegible] swamping resistor, R19, is [illegible] to provide adequate [illegible] and [illegible] efficiency in the amplifier.

RF chokes L15 and L16 should be glued to the pc board by means of silicone rubber adhesive. The other end [illegible] wound with [illegible] and are supported adequately by their leads.

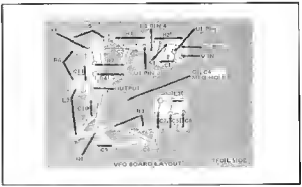


Fig. 4 — Full-size template for the VFO board.

Q8 is the only transistor in the unit that requires a heat sink. It is shown with the sink removed in Fig. [illegible] for clarity. The heat sink in the author's unit is a Thermalloy type 2201A with two 1-inch-square pieces of aluminum bolted to it. It can be observed in photographs. The sink is [illegible] and the key must be depressed for a good while before the assembly becomes warm. Almost any of the finned clip-on sinks that are readily available should work well.

The power amplifier is keyed on by applying [illegible] excitation to the input rather than by keying V_{cc}. When the first prototype was being built, it was found that trying to key [illegible] caused clipping, particularly when the current through them exceeded 300 mA. Only by keying the driver was this problem eliminated.

There is no danger of thermal runaway with this arrangement because the base of Q8 is tied to the emitter through L15. Also the leakage current consumed by Q8 when it is not in use is negligible. It is on the order of a few microamperes.

This shows the inside of the transceiver as viewed from the rear.

As a matter of interest, the transmitter was operationally checked at 10° F increments between –40° F and +140° F. The output was stable and no amplitude change could be observed. Rf power output is typically 5 watts for 8 watts input.

Construction

The receiver, driver, and power amplifier boards all measure 3-1/2 × 2-1/8 inches and the VFO board is 2 1/8 inches square. Layout of the boards is not critical and most any convenient packaging arrangement may be used. All rf wiring is done with RG-174/U. Extra solder lands for interconnections between the VFO, driver, and receiver boards are provided on the receiver board.

The unit is housed in a homemade aluminum box measuring 2-3/8 inches high by 4 inches wide, and 6-3/4 inches deep including the 5/8-inch front overhang of the top cover. The chassis is finished in Golden Harvest Shadow epoxy appliance enamel and the top cover is painted with a dark brown wrinkle finish. Aramids lettering is protected with a coat of clear acrylic spray. Silicon rubber bumpers are used as feet to complete the cabinet.

The VFO output was brought out to a jack on the back panel so that a frequency counter could be used for a digital frequency readout when operating is home. The whole station, including the key, earphone, NiCad battery pack, and a 40-meter dipole can be carried in an ordinary lunch bucket.

Alignment

Alignment of the VFO is accomplished by monitoring its output frequency with a frequency counter or calibrated receiver. Tuning the output frequency to 7.0 MHz by adjusting C3 with C4 while they are fully meshed is also done. The receiver is aligned by tuning in a station at 7.075 MHz and adjusting C13 for maximum headphone volume. For transmitting alignment, a dummy load with rf detector as shown in Fig. 8 should be used.

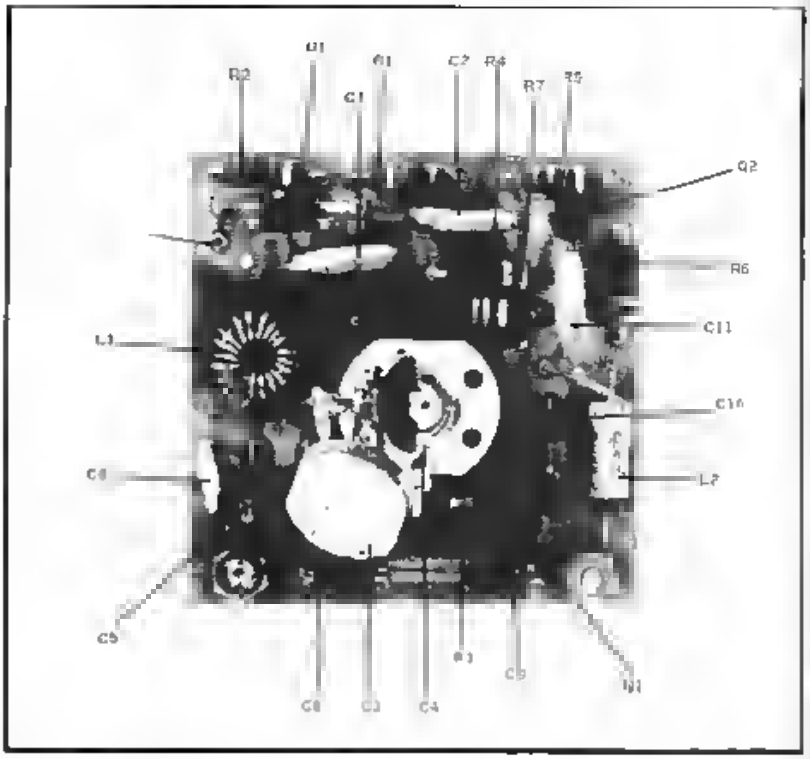


Fig 5 - Parts placement for the VFO board.

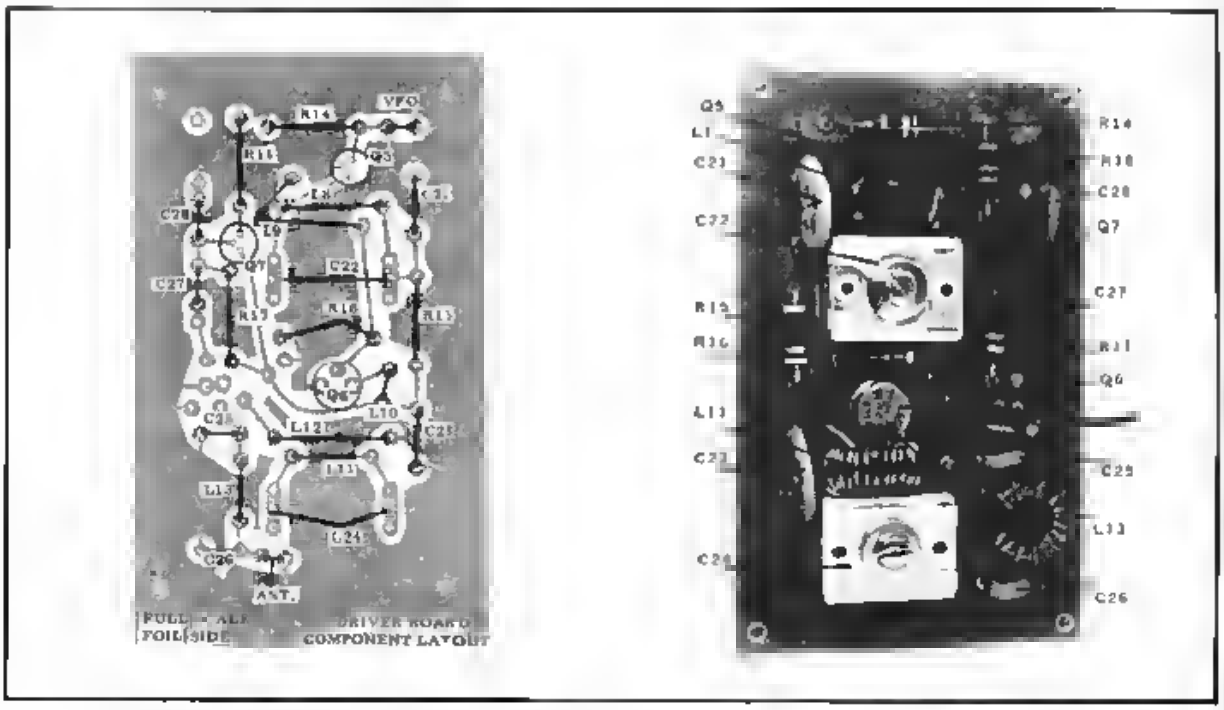


Fig 8 — Full-size template and parts placement for the driver board.

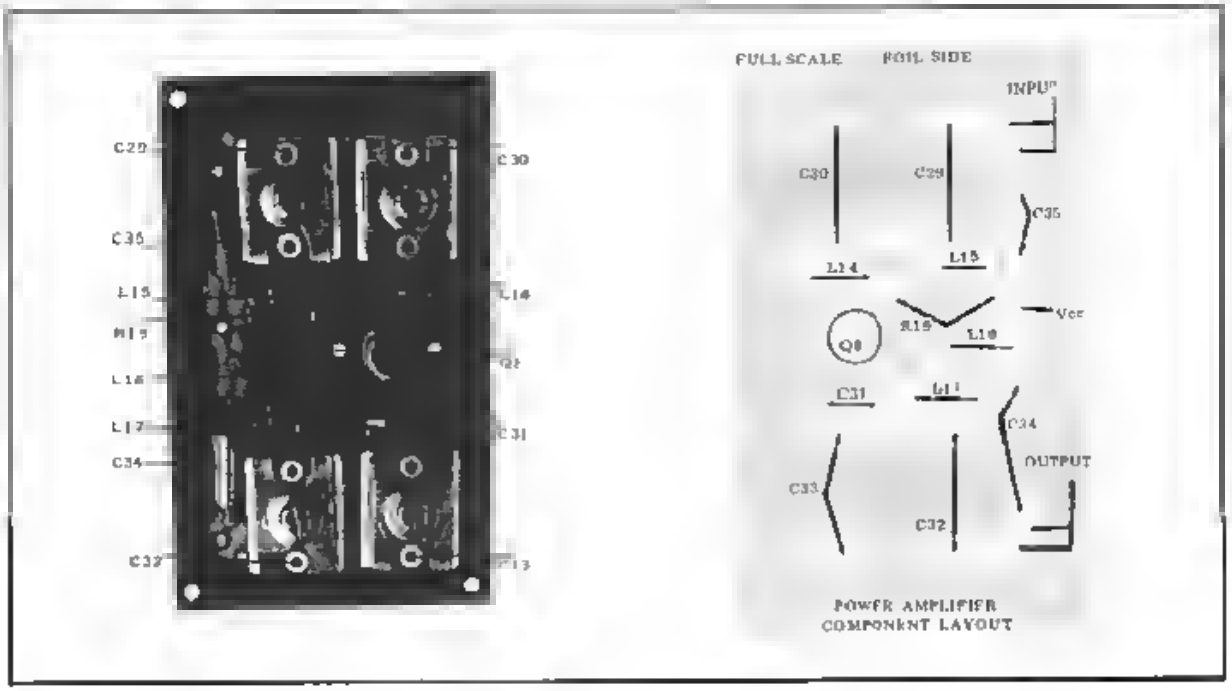


Fig 7 – Full-size template and parts placement for the power amplifier board.

Turn the adjusting screws of C22, C21, C29, C30, C32 and C33 to maximum clockwise positions. With the load connected to the antenna terminals S3 set to low power, and Vcc set at 10 V dc, depress the key and adjust C22 and C24 for maximum output. Then increase Vcc to 12 V dc and repeat the adjustment. The tuning should be smooth and regular. Next set Vcc back to 10 V dc, S3 to high power and adjust C29, C30, C32 and C33 for maximum output. They interact so you will find it necessary to go back over them a few times until no further increase in output can be obtained. Increase Vcc to 12 V dc and repeat the procedure. The repetition should require very little retuning and should cause the output to vary smoothly with no sudden variations.

After tune-up, a battery current-drain check should yield the following values with Vcc at 13.6 V dc:

Receive mode	20 mA
Transmit mode (low)	100 mA
Transmit mode (high)	750 mA

The author wishes to thank the staff, faculty, and students at the Mankato Area Vocational-Technical Institute who provided assistance on this project.

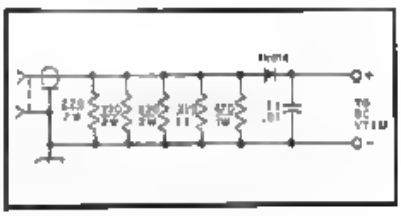


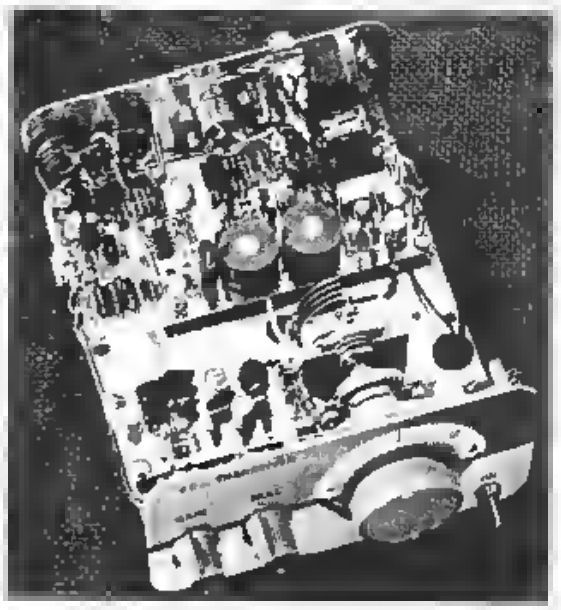
Fig 8 – Dummy load and rectifier.

From October 1985 QST, p 16.

Better Ears for the MAVTI-40 Transceiver

A transceiver need not be a complicated building project. Try QRP—and instead of tackling a superhet receiver, take the direct approach!

By Paul Kranz, W1CFI
26 Mettscomett Path
Harvard, MA 01451



Although this article concentrates primarily on the redesign of the MAVTI 40 receiver section, there's enough information here to permit you to build a complete 40-meter QRP (low-power) transceiver. PC boards and parts kits are available to make your job even easier.

Direct-conversion (D-C) receivers are easier and less costly to build than their superheterodyne cousins, and assembling a D-C receiver is an educational and rewarding project. Although D-C receivers have some performance shortcomings, the receiver described here eliminates some of them. This receiver will reject AM broadcast interference to the level of inaudibility. It also provides a narrow-bandwidth filter for CW reception and a tunable notch filter. Modifications for improving the stability and wave shaping of the original MAVTI-40 transmitter are provided, as is a TR switch.

A Club Transceiver Project

In 1979, the Hewlett-Packard Amateur Radio Club in Andover, Massachusetts, began a Novice class with five prospective radio amateurs. Since none of them had any equipment, we decided that a simple transceiver construction project might solve this problem as well as offer some experience working with hardware. A search of back issues of Amateur Radio magazines turned up one transceiver design that seemed to offer many advantages over other designs. This transceiver, the MAVTI-40, had originally been designed and constructed as a radio club project, and several had been built.[1] This suggested that the 5-W-output, 40-meter tranrceiver

[1]Notes appear at end of article

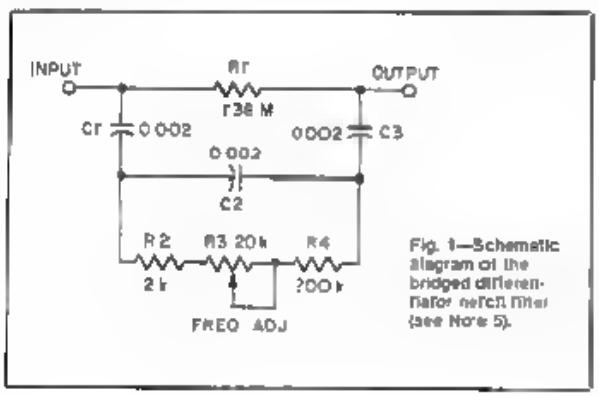


Fig. 1—Schematic diagram of the bridged differentiator notch filter (see Note 5).

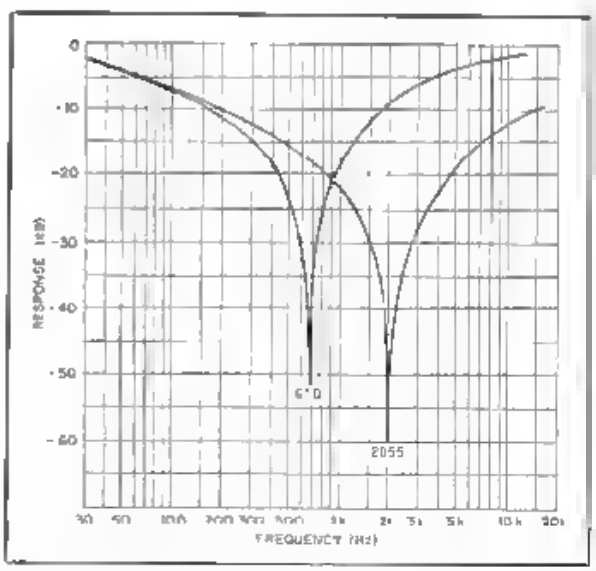


Fig. 2—Frequency response of the notch filter of Fig. 1 with the potentiometer set at each end

should be capable of being duplicated easily without the problems associated with many one-of-a-kind designs. Further, a PC-board negative was available from the author, making the construction repeatable and reliable.

Forty meters is a good band for beginners because it has an active Novice segment during daylight and evening hours. Also, the band offers good DX and QRP activity in the General- and higher-class portions of the band, and that encourages license upgrading. One disadvantage of 40-meter operation is the evening hour AM-broadcast interference.

Five MAVTI-40 transceivers were built using a variety of construction techniques. Although the transceivers performed reasonably well, they all exhibited occasional instability in the transmitter and receiver sections. One of the units has been in use at my station for five years, and has served as a test bed for many experiments and subsequent improvements to the original design. Eventually, the instability problems were solved and the transceiver has provided many enjoyable contacts.

Receiver Improvements

The original receiver was difficult to use at night because it detected AM broadcast stations that resided more than 100 kHz above the usual 7040-kHz QRP operating frequency. This problem became increasingly worse as the sunspot activity declined. Several initial modifications, including the use of different mixers and additional input filtering, were tried without success. An examination of Amateur Radio magazine articles turned up some 40-meter transceiver designs that addressed the AM-detector problem.[2,3] These articles offered the inspiration needed to attempt a redesign of the original receiver.

Mixing Schemes

Measurements made at my location revealed broadband signals of 100-mV P-P at the feedpoint of a dipole antenna. These signals would need to be removed before they reached the mixer. AM broadcast stations at 7.2 MHz produced 8-mV P-P on a 50-ohm load connected to the antenna, while the strongest CW signals measured 50-µV P-P. AM-detection comparisons were made using an HP-3585A spectrum analyzer coupled to the MAVTI-40 MOSFET mixer, a harmonic detector (see Note 2) and doubly balanced mixers. These measurements were made by injecting a 50% amplitude-modulated signal into the mixer RF input while measuring the detected AM signal with the analyzer. The frequency of the AM input signal was chosen to be 100 kHz above the mixer local oscillator (LO) to simulate actual 40-meter operating conditions. The detected AM signal is the actual audio modulation (baseband). The result is expressed as a decibel ratio between this audio signal and the mixer output when the LO is tuned to receive the AM signal. The MAVTI-40 mixer was able to reject this AM signal by only 35 dB. The harmonic detector (with the LO operating at half the RF input frequency) rejected the unwanted AM signal by 60 dB. A doubly balanced diode mixer was the best performer,

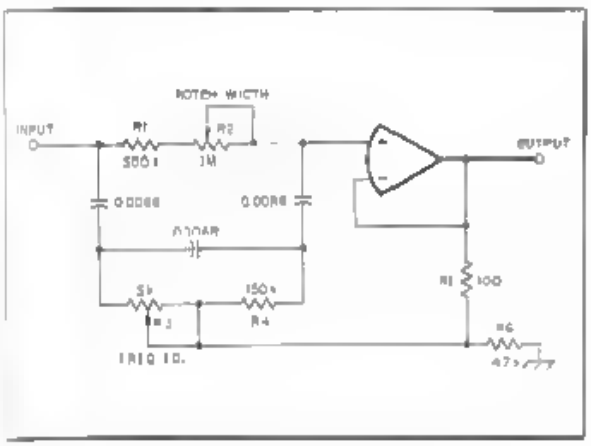


Fig 3—Schematic diagram of the active notch filter. R3 controls the notch frequency. This filter has a gain of 0 dB.

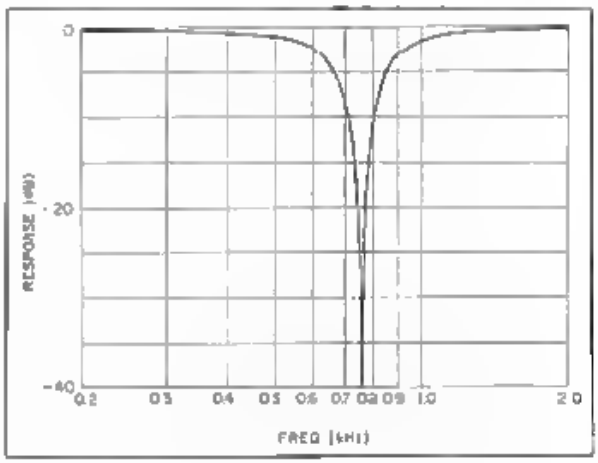


Fig 4—Frequency response of the tunable active notch filter of Fig 3.

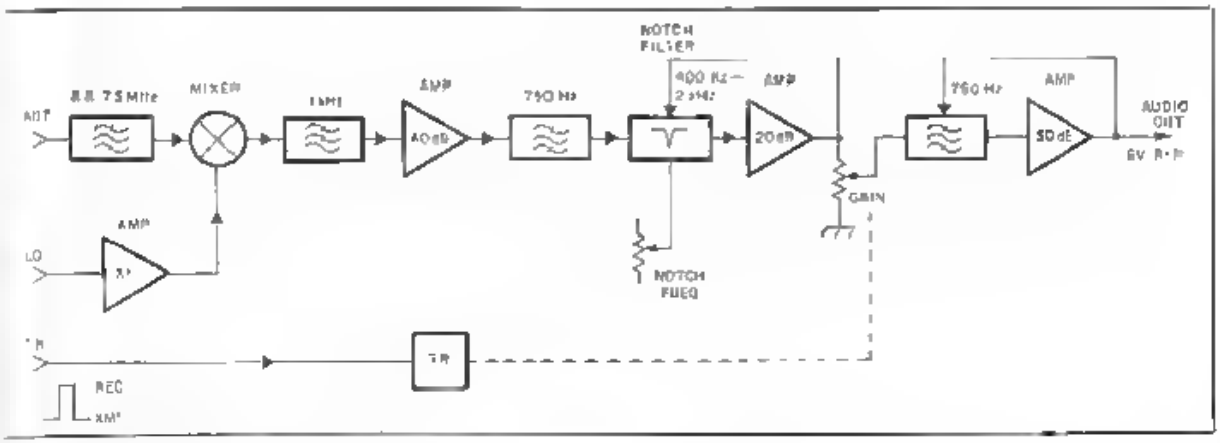


Fig 5—Block diagram of the new D-C receiver.

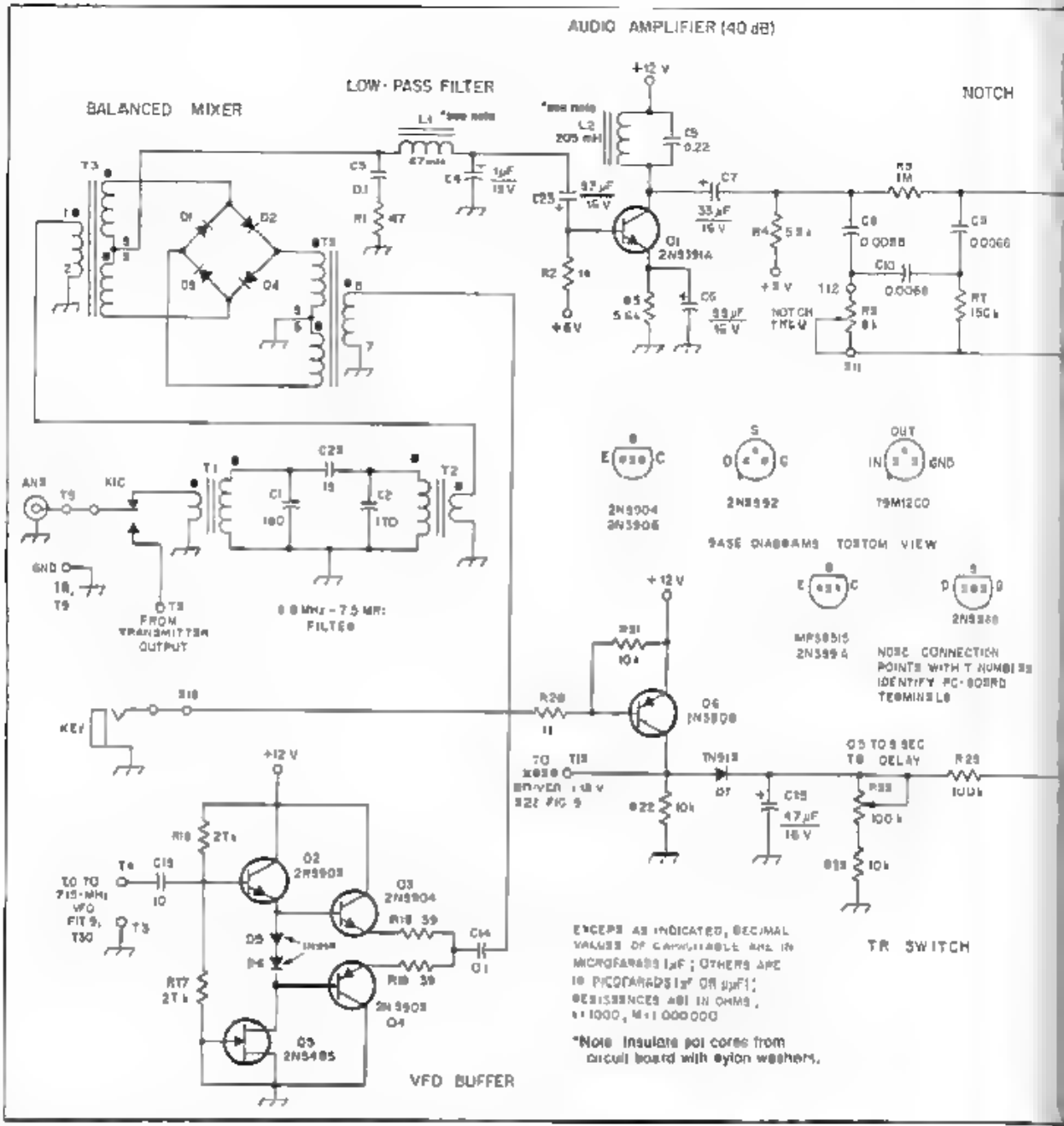


Fig 8—Schematic diagram of the D-C receiver including the TR switch. Note: Equivalent parts may be substituted. Unless otherwise specified enameled wire is used for winding inductors.

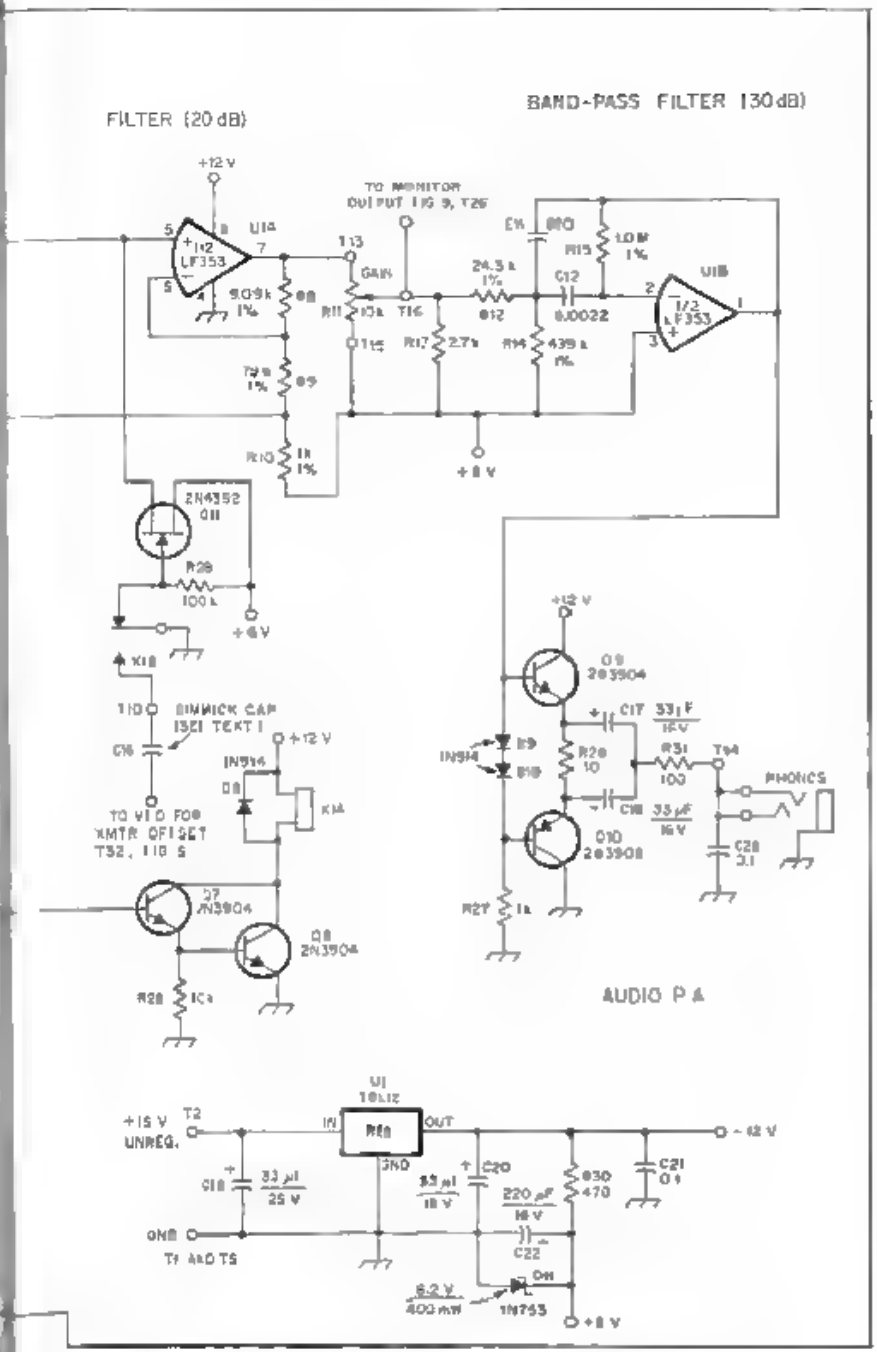
D1-D4, incl—HP2800 hot carrier diode or part of U3 (see Text).
D5-D10, incl—1N914 or 1N4148.
D11—1N753 6-V, 0.4 W Zener diode.
K1—12 V, DPDT (Radio Shack 275-213).
L1—113 turns no. 26 on Amidon pot core PC 2213-77.
L2—237 turns no. 30 on Amidon pot core PC 2213-77.
Q1—2N3391A, MPS6515.
Q2, Q3, Q7-Q9, incl—2N3904.
Q4, Q6, Q10—2N3906.
Q5—2N5485 FET.
Q11—2N4392.
T1, T2—Primary, 4 turns no. 30 on Amidon

boasting a 73-dB rejection ratio.

Compared to the harmonic detector, the doubly balanced mixer has the additional advantage of being insensitive to the LO waveshape. In fact, this mixer is most efficient when driven by a square wave. The mixer diodes are used as switches and, as such, do not provide mixing by virtue of their nonlinear transfer curves as they do in the harmonic mixer. A disadvantage of the doubly balanced diode mixer is the amount of LO power required, typically 7 dBm.

Noise Figure

Although the atmospheric noise in the 40-meter band is not so low that a low-noise-figure receiver is required, an attempt was made to keep the receiver noise to a reasonable level. Since the noise figure will never be lower than the mixer conversion loss (6 to 8 dB), the remaining amplifiers serve only to make the noise figure worse. Atmospheric noise in a quiet location contained in a 200-Hz bandwidth on 40 meters



T25 2 cores, secondary 28 turns no. 30.
T3, T4—15 turns no. 30 bifilar wound on Amidon FT-50-43 core or part of U3 (see text).
U1—LF353N dual FET op amp.
U2—78M12CG, 12-V, 500-mA, 3-terminal regulator.
U3—Mini-Circuits SBL-1 doubly balanced diode mixer (see text).

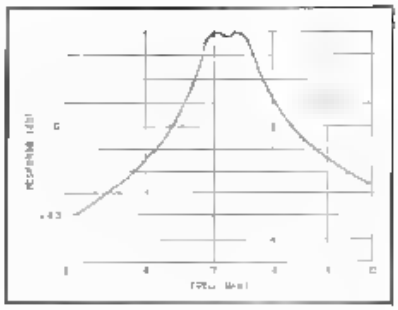


Fig 7—Input band-pass filter frequency response.

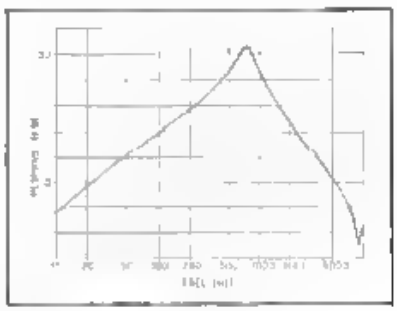


Fig 8—Band-pass characteristics of the D-C receiver (see text).

has been shown to be approximately 0.4-µV RMS.[4] This amount of noise would require a receiver noise figure of 20 dB (10 dB S+N/N) where the receiver noise would be just equal to the atmospheric noise.

Single-Signal Reception

One of the major shortcomings of D-C receivers is their lack of single-signal reception. When a CW station is tuned in, it can be heard equally well when the VFO is tuned above or below the zero-beat frequency. This characteristic has the effect of doubling the number of stations falling in the receiver audio passband, compared to what a superheterodyne receiver would produce. Some solutions to this problem add complexity to the D-C receiver and result in a component count that differs little from that of a superheterodyne receiver. A tunable notch filter can be used to null out an offending signal and goes a long way toward solving the single-signal reception problem.

There are many notch-filter designs described in the literature; however, one design offers notch-frequency adjustment with only one potentiometer.[5] This bridged-differentiator circuit is shown in Fig. 1, and a plot of its response is given in Fig. 2. The main problem with this design is the width of the notch at frequencies above and below the notch frequency. The addition of feedback from an op amp solves this problem and provides a notch depth of 40 dB. The resulting circuit is shown in Fig. 3, and a plot of its response is in Fig. 4. In Fig. 3, R3 adjusts the notch frequency, while R5 is used to adjust the notch width, or Q. R2 maximizes the notch depth at a given frequency. Test results of the circuit show a tunable range of 400 Hz to 2 kHz, and a notch depth of 30 to 40 dB for the component values shown. This notch depth is adequate since deeper, higher-Q notches do not take into account the finite bandwidth of CW so the operator will still be able to hear key-click-like sounds from the offending station.

The best solution to the problems experienced by the MAVTI-40 receiver seemed to be to design a completely new

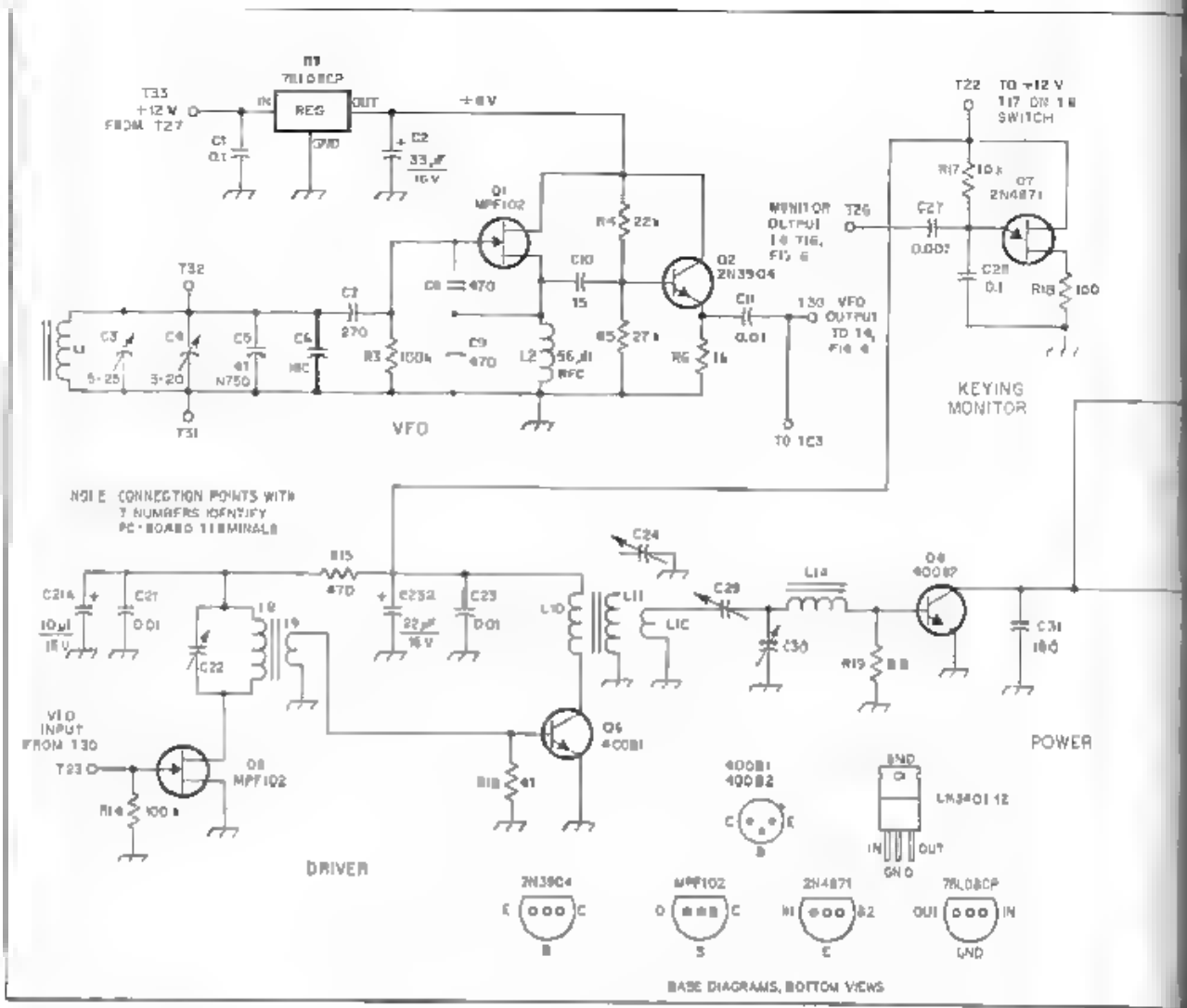


Fig. 6. Schematic diagram of the VFO and modified MAVTI-40 transmitter. Components are numbered to agree with those of the original MAVTI-40 VFO and transmitter. Therefore, designators Q3 and Q4 are missing. Note: Equivalent parts may be substituted. Unless otherwise specified, enameled wire is used for inductor windings.

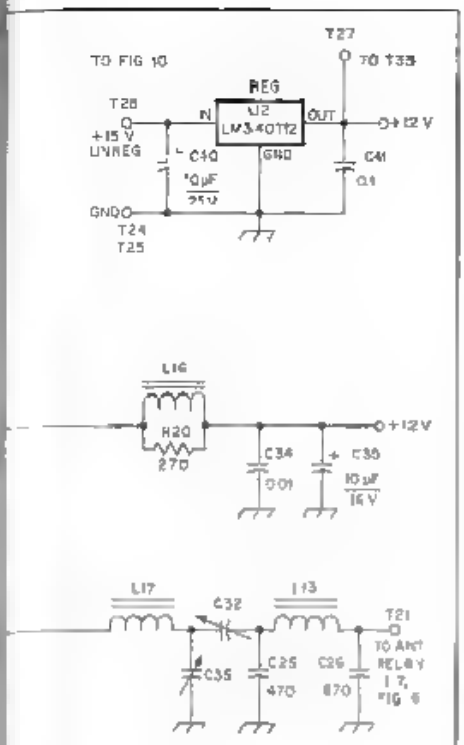
C3—5-25 pF trimmer
C4—3-20 pF air variable (Johnson 160-110-5)
C5—47 pF N750 temp compensating capacitor
C22—10-180 pF compression trimmer (ARCO 463)
C24, C28, C30, C32, C33—75-480 pF compression trimmer (ARCO 466)
L1—1.4 µH, 18 turns no. 24 enam. wire on Amidon T37-2 core
L2—56 µH molded RF choke.
L8—34 turns no. 26 on Amidon T50-2 core
L9—3 turns no. 24 on L8
L10—7 turns no. 24 on L11
L11—22 turns no. 22 on Amidon T50-2 core
L12—4 turns no. 26 on L11
L13—14 turns no. 20 on Amidon T50-2 core
L16, L17—20 turns no. 24 on Amidon T37-2 core
L18—65 turns no. 32 on Amidon T37-2 core
Q1, Q5—MPF102, 2N4416 or 2N5486 FET
Q2—2N3904
Q3—400B1 or MRF 8003 RF power transistor
Q4—2N4871 enhancement transistor
Q6—400B2 or MRF 8004 RF power transistor
U1—78L08CP, 8-V, 100-mA, 3-terminal regulator
U2—LM340T12, 12-V, 1-A regulator

receiver incorporating these improvements. A doubly balanced diode mixer solves the AM-detection problem and provides good immunity to third-order intermodulation distortion. The active audio filtering offers a 200 Hz bandwidth for CW reception. A tunable notch filter helps reduce interference from adjacent signals and the undesired audio image frequency common to D-C receivers. A block diagram of the new receiver is shown in Fig. 5, and its performance figures are given in Table 1.

Receiver Circuit Description

The receiver schematic diagram is shown in Fig. 6. Signals arriving from the antenna enter the receiver through the TR relay contacts K1C and the input band-pass filter (T1, T2, C1, C2, C24). The filter has a passband ripple of about 3 dB from 6.8 to 7.5 MHz, its frequency response curve is presented in Fig. 7.

A Mini-Circuits Labs SBL-1 doubly balanced diode mixer is used in my receiver.[6] Any doubly balanced diode mixer may be used including a "homebrewed" version.[7] The LO drive is supplied by a buffer amplifier consisting of Q2-Q5. The mixer output is terminated for RF signals by C3 and R1. Audio output from the mixer is filtered and impedance matched to the input of Q1 by the low-pass filter consisting of L1 and C4. Because of its im-



EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF OR µµF). RESISTANCES ARE IN OHMS; K=1000, M=1000 000

Table 1
Receiver Performance Specifications

Sensitivity (10 dB S + N/N)	0.4 µV (−115 dBm)
Bandwidth (audio IF of 750 Hz)	200 Hz
Gain	90 dB
Noise figure	20 dB
Third-order intercept	+13 dBm
Dynamic range	96 dB
AM detection	−70 dB
Notch depth	40 dB
Notch frequency	400 Hz to 2 kHz

pedance transformation, this filter has a voltage gain of 6 dB, which helps to make up for the mixer conversion loss.

The first receiver amplifier stage, Q1, is designed for low noise while providing a gain of 40 dB in a 200-Hz bandwidth. This bandwidth is controlled by the Q of L2 and the resistance of R4.

U1A acts as a 20-dB-gain tunable notch filter. The notch frequency is adjusted from 400 Hz to 2 kHz by R6. Notch-filter Q is controlled by R9 and R10, and seems to be adequate for CW. Since the notch depth changes from 30 to 40 dB as the notch frequency is varied, R5 can be selected for best notch depth at your preferred frequency. The value of R5 will vary with the tolerance and matching of C8-C10.

The band pass filter, U1B, provides a gain of 30 dB at 750 Hz with a bandwidth of 200 Hz. This brings the total receiver gain to 90 dB. Fig. 8 shows the band-pass characteristic of the complete receiver from the mixer output through the band-pass filter. The notch filter has been set to a high frequency in order to remove the notch from the plot.

The output of the band-pass filter is buffered by Q9 and Q10, which provide sufficient power gain to drive a pair of low-impedance headphones, such as those used with a personal stereo radio. Because of the large amount of gain (90 dB) at 750 Hz, it is not possible to use this amplifier to drive a speaker and still maintain stable operation at full gain.

Fig. 6 also shows a TR switch, Q6-Q8. Keyed power for the original MAVTI-40 driver PC board is derived from the collector of Q6. A turn-off delay for the TR relay, K1, is produced by C15 and R23. The delay is adjustable from 0.5 to 5 seconds by adjustment of R23. One pair of K1 contacts (K1C) switches the antenna between the receiver and transmitter. Another contact set (K1B) turns off the receiver mute switch (Q11, R29) during receive periods. These contacts also provide a convenient way to shift the VFO frequency down by 750 Hz during transmission. This is accomplished by grounding a gimmick capacitor (C16) connected between normally open relay contact (K1B) and the VFO tuning capacitor. (The gimmick capacitor is a small-value capacitor made by twisting together two pieces of insulated wire.) The capacitor is trimmed to the correct value by cutting away small portions of the wire while measuring the frequency shift with a frequency counter or another receiver.

Transmitter Improvements

While I was adjusting the mica compression trimmers in the original transmitter section, the RF output across a 50-ohm dummy load jumped suddenly to maximum output. This behavior suggested that the transmitter section was oscillating at or near the VFO frequency. The transmitter instability was solved by making three minor changes to the original circuit.

Fig. 9 shows the schematic diagram of the VFO and modified MAVTI-40 transmitter, including the corrections. In the original article R16 is lowered to 47 ohms. Next, a 270-ohm resistor is added in parallel with L16. Finally, L15 is removed from the base of Q8 since the base resistor, R19 (6.8 ohms), provides adequate stability for this power amplifier. The transmitter section now tunes up smoothly to a 4-W output level, and no instability has been observed. Note that this solution to transmitter instability worked well on my transmitter, and some variation from one transmitter to another may require minor changes.

The keyed transmitter output of the original MAVTI-40 has a square-wave envelope since no attempt was made to shape this waveform. I've added a 10-µF capacitor in parallel with C21 and a 22-µF capacitor in parallel with C23 to provide output waveform rise and fall times of approximately 5 ms. I have received many compliments concerning the clean sounding QRP signal from this transceiver.

Construction

The receiver, VFO and transmitter sections of the transceiver are constructed on three PC boards, which are mounted inside an LMB CO-3 cabinet. The PC boards are double sided with the top side of each board serving as a ground plane; the boards have plated-through holes.[1]

I use an external, unregulated supply (Fig. 10) to power the transceiver. It's probably best not to include this power supply inside the transceiver cabinet since hum pick up could become a problem.[1] Voltage regulation is provided for on the receiver board by a three-terminal regulator. The 6-V supply is derived from the regulated 12-V line by using a 470-ohm resistor in series with a 6.2-V Zener diode (D11).

I fashioned a tuning dial by attaching a clear plastic disc to the mounting plate of a Jackson Brothers 30:1 reduction drive. Calibration marks are made by applying

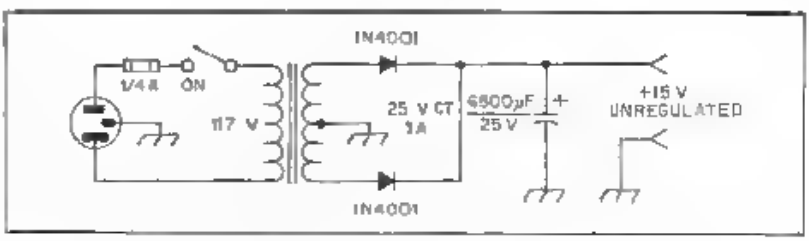


Fig. 10—Schematic diagram of the unregulated power supply.

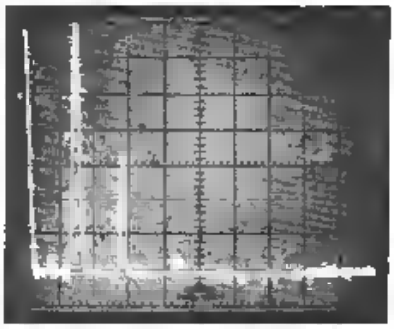


Fig 11—Spectral display of the transmitter output. Vertical divisions are each 10 dB, horizontal divisions are each 5 MHz. Output power is 4 W at 7 MHz. All spurious emissions are at least 32 dB below peak fundamental output power.

dry transfers to a translucent piece of plastic film cemented to the plastic dial plate with rubber cement. Another piece of clear plastic is used for a dial window, and a hair-line indicator was produced by scratching the plastic with a scribe. The window is cemented to the inside of the front panel behind the dial cut-out and in front of the dial plate. Back lighting for the dial is provided by two small, colored lamps.

L1 and L2 are mounted to the PC board using the plastic mounting screws supplied with the pot cores. The gimmick capacitor, C16, is made from two pieces of no. 22 insulated, solid-copper wire twisted together over a length of 1 inch. C16 is connected between the VFO tuning capacitor and terminal T10.

The transmitter-board inductors are wired to the board in two different ways. L8, L9, L10 and L12 have each winding connected to the board at opposite sides of the toroid. All other inductors have their windings connected to the board on the same side of the inductor. The mica compression trimmer capacitors are mounted by soldering a U-shaped piece of no. 22 bare wire to each solder lug on the capacitor. The bus wire is then inserted into the two holes in the PC board. The 12-V regulator uses the PC board mounting stud as its heat sink. A heat sink must be used with Q8.

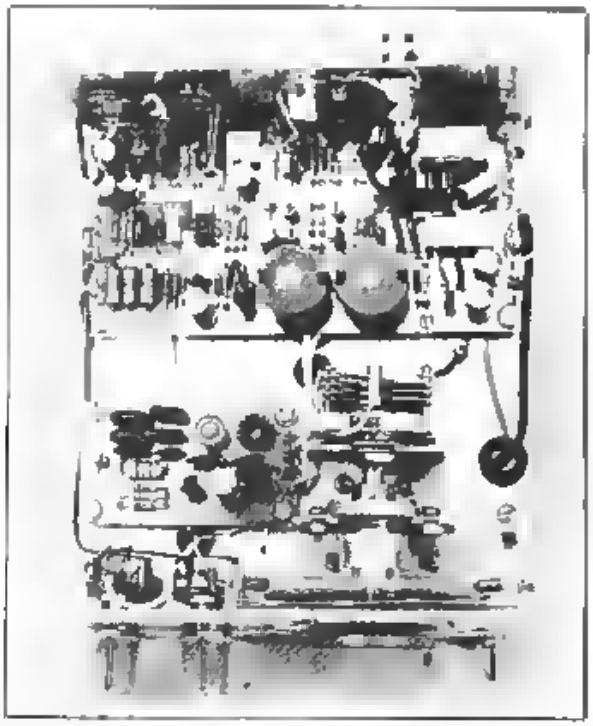
All boards should be tested (refer to the next section) before they are mounted in the transceiver. The PC boards are interconnected with unshielded wire in all cases except for the antenna-to-TR-switch and TR-switch-to-transmitter connections. RG-174 miniature coaxial cable is used for the latter connections. To avoid the possibility of creating unwanted oscillations, interconnecting wires should not run beneath the receiver board. PC-board terminals are used on my transceiver boards, but the wires can be soldered directly to the PC board. Notes concerning interconnection of the boards appear adjacent to each terminal in the transmitter schematic diagram, Fig. 9.

Initial Tests and Calibration

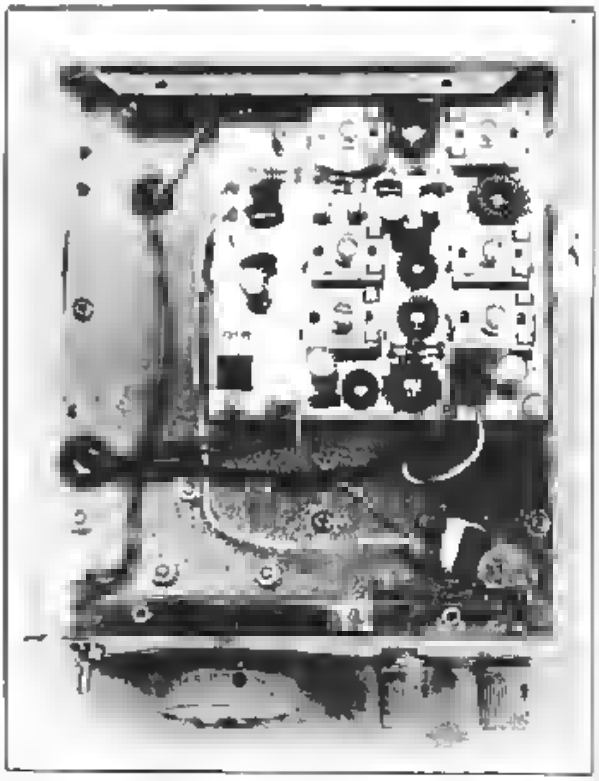
Because the VFO is needed to drive the receiver and transmitter boards, check it first. You may operate the VFO directly from the unregulated 15-V supply during these tests. See that U1 is supplying 8-V dc output, and that an RF signal output of approximately 4-V P-P is present at T30. For the moment, that's all for the VFO; its calibration will be done later.

With 15-V dc applied to the receiver board, check that U2 provides 12-V dc output, and approximately 6-V dc is present at T15. Pins 1 and 7 of U1 should be at the same potential as T15. Connect a pair of headphones between T14 and ground. Short T13 and T16 together, while noise should be heard in the phones. Shorting T18 to ground should close K1, and adjusting R23 should vary the release delay from approximately 0.5 to 5 seconds.

Tune up the transmitter as follows. Connect a 5-W dummy load between T21 and



A top inside view of the author's transceiver. The VFO PC board is at the front left, behind the GAIN and NULL potentiometers. The receiver board is at the rear of the unit. At the front right-hand side of the board is the doubly balanced mixer module. Almost directly behind it, near the rear panel, is the T/R relay. The two cylindrical objects at the front center of the board are L1 and L2.



The transceiver viewed from the bottom. In this prototype the RF output transistor has an elongated heat sink that runs parallel to the right side of the board.

T25. Set all mica compression trimmer capacitors for maximum capacitance (fully closed). Key the transceiver and see that 12-V dc is present at T22. Adjust C22 for maximum RF output across R16. Then adjust C24 for maximum RF voltage across L11. Set C29 and C30 for maximum RF voltage across R19. Last, adjust C32 and C33 for maximum output across the dummy load. Since C29, C30, C32 and C33 adjustments interact, the process will have to be repeated several times. During the final stages of tune-up, the trimmer-capacitor adjustments should provide smooth amplitude variations with no sudden jumps apparent. Monitor the temperature of Q8 closely during transmitter tune-up.

Adjust C3 to have the VFO cover the desired frequency range, and set the dial calibration. C5 provides temperature compensation. No noticeable drift should occur after an initial warm-up period of about 10 minutes.

The VFO offset during transmit is set by trimming the length of the gimmick capacitor, C16. Trim C16 to provide a downward VFO frequency shift of about 750 Hz when the transmitter is keyed.

Operation and Comments

Since the transmit frequency is shifted below the receive frequency, it is necessary to tune the receiver so the VFO frequency is above that of the received station. When the transmitter is keyed, the VFO frequency shifts down by 750 Hz and falls on the zero-beat frequency.

The transceiver has been in use for several months, and the improved receiver performance makes the redesign effort worthwhile. There is absolutely no audible amplitude modulation from the high-power 40-meter broadcast stations. The bandwidth of the receiver is adequate for CW reception, and no audio distortion or ringing is evident. In fact, the audio signal has good tone quality when personal stereo headphones are used with the receiver. The notch filter has proved useful, however, it is not a complete substitute for single-signal reception. When the transceiver is used to work other QRP stations, it is helpful to have a low-noise receiver since the received signals can be just above the 40-meter-band noise during the daylight hours. Get out your soldering iron and try your hand at building the receiver or the entire transceiver. I'm sure you'll be glad you did!

Acknowledgment

I'd like to thank Jim Conrad, N1GW, of Hewlett Packard, for his suggestions concerning the design of the receiver RF section and for his help with AM-detection measurements.

Notes

[1]D. K. Seaman, "The MAVTI-40," *QST*, June and July 1975. Also see Feedback, *QST* Oct 1975, p. 71.

[2]J. L. Kath, "40-Meter Transceiver for Low Power Operation," *Ham Radio*, April 1980.

[3]R. W. Lewellen, "An Optimized QRP Transceiver," *QST*, Aug. 1980.

[4]J. A. Dyer, "High Frequency Receiver Performance," *Ham Radio*, Feb. 1984.

[5]C. Hall, "Tunable RC Notch," *Ham Radio*, Sept. 1975.

[6]Min-Circuits Labs, P.O. Box 166, Brooklyn, NY 11235, tel. 718 934-4500.

[7]C. Hutchinson, ed., *The 1985 ARRL Handbook for the Radio Amateur* (Newington: ARRL, 1984), p. 12-19.

[8]Circuit boards and parts kits are available from Radiokit, P.O. Box 411, Greenville, NH 03048, tel. 603-878-1033. PC-board templates and parts overlays are available (note: this is a *double-sided* board) from ARRL HQ for $2 and a business-sized SASE. Address your correspondence to the Technical Department Secretary, and identify your request as MAVTI Updates.

[9]Editor's Note: See D. DeMaw, "Plug-In Wall Transformers—A Super Bargain," *QST*, June 1985. On page 37, Doug relates some of his experiences with DC receivers and hum pickup.]

For updated supplier addresses, see ARRL Parts Supplier List in Chapter 2.

From December 1989 QST, p 18:

A QRP SSB/CW Transceiver for 14 MHz

Part 1: Exotic circuitry and hard-to-find components aren't necessary if you want to build excellent performance into a home-brew SSB/CW transceiver: Careful design is the key.

By Wes Hayward, W7ZOI
7700 SW Daniella Ave
Beaverton, OR 97005

It's hard to justify the construction of a complete SSB/CW transceiver in this "modern" era of readily available commercial equipment. The popular, multiband MF/HF transceivers offer excellent performance, often at a reasonable cost. Still, I feel a twinge of guilt when I use them. They offer nothing of the feeling of exploration that I've grown to expect from Amateur Radio.

The rig described here is not a copy of the usual "appliance." I've used the project as a vehicle to investigate alternative circuits and a block diagram that departs from the traditional. The circuit is simple and modular, with flexibility that allows for later changes.

I present this rig in order to encourage other home-brew enthusiasts to give QRP SSB a try. I'll not dwell on the standard circuits that are already covered in *Solid-State Design* or in *The ARRL Handbook*.[1,2] Rather, I'll emphasize only those circuits that depart from the traditional. This is intended to be an idea article rather than a construction piece. There are no circuit boards or patterns available for this rig. All construction was done using "ugly" methods.[3]

System Architecture

The filter method was chosen for this transceiver. While that is generally considered to be "the only choice," phasing methods should not be overlooked for an experimental transceiver.[4] The block diagram is shown in Fig 1.

The traditional filter transceiver shares one or more crystal filters between the receive and transmit modes. I wanted to avoid the compromises and complexities of filter switching, so I decided to use separate filters for each function. The transmit and receive modules can then be used for completely independent operation. This might be especially interesting for use with, for example, a VHF/UHF station for OSCAR communications.

Commercial crystal filters from my junk box were used in this project. They are all 9-MHz circuits that are, fortunately, well matched to each other. A 5-MHz local oscillator drives both the receiver and transmitter mixers. Budget-minded builders may elect to build their own filters.[5,6]

The Receiver

The receiver is very much like the Progressive Receiver that's been in *The ARRL Handbook* for several years.[7] The front end and VFO are presented in Fig 2. I initially used a VFO variable capacitor with a vernier drive mechanism. Problems occurred with the mounting, however. The VFO was rebuilt without a vernier. Instead, two capacitors were used. One (C1, BANDSET) tunes the entire band, while the other (C2) is a bandspread control with a total range of only 25 kHz. This scheme seems to be practical for a simple transceiver.

The receiver begins with a doubly tuned preselector and a diode-ring mixer (U1, a Mini-Circuits SBL-1). This is followed by a bipolar transistor (Q3, an NEC99532) in a negative-feedback IF amplifier. A ferrite transformer (T4) matches the IF amplifier to the receiver crystal filter (FL1) as shown in Fig 3. The filter I used is similar to the KVG XF-9B. The less-expensive KVG XF-9A was tried in this application and was found wanting for stop-band attenuation.

The crystal filter drives an MC1350P IF amplifier (U2) and a diode-ring product detector (U3, an SBL-1). I would discourage a builder from departing from a diode-ring detector. An NE602 detector was tried, but suffered from severe in-band intermodulation distortion.

The BFO signal is low-pass filtered before driving the detector. A reduced-voltage sample of the BFO energy is routed to the transmit balanced modulator (to be described in Part 2 of this article). Care was taken to extract the sample from a point away from the detector. (The diode-ring detector clips the BFO waveform; clipped carrier-oscillator drive for the balanced modulator is undesirable.)

The audio amplifier (Q6-Q8 and U4) is standard. However, the audio-derived AGC system departs from the usual. U5A (one section of an LM324) amplifies the audio IC a level suitable for

[1]Notes appear at end of article

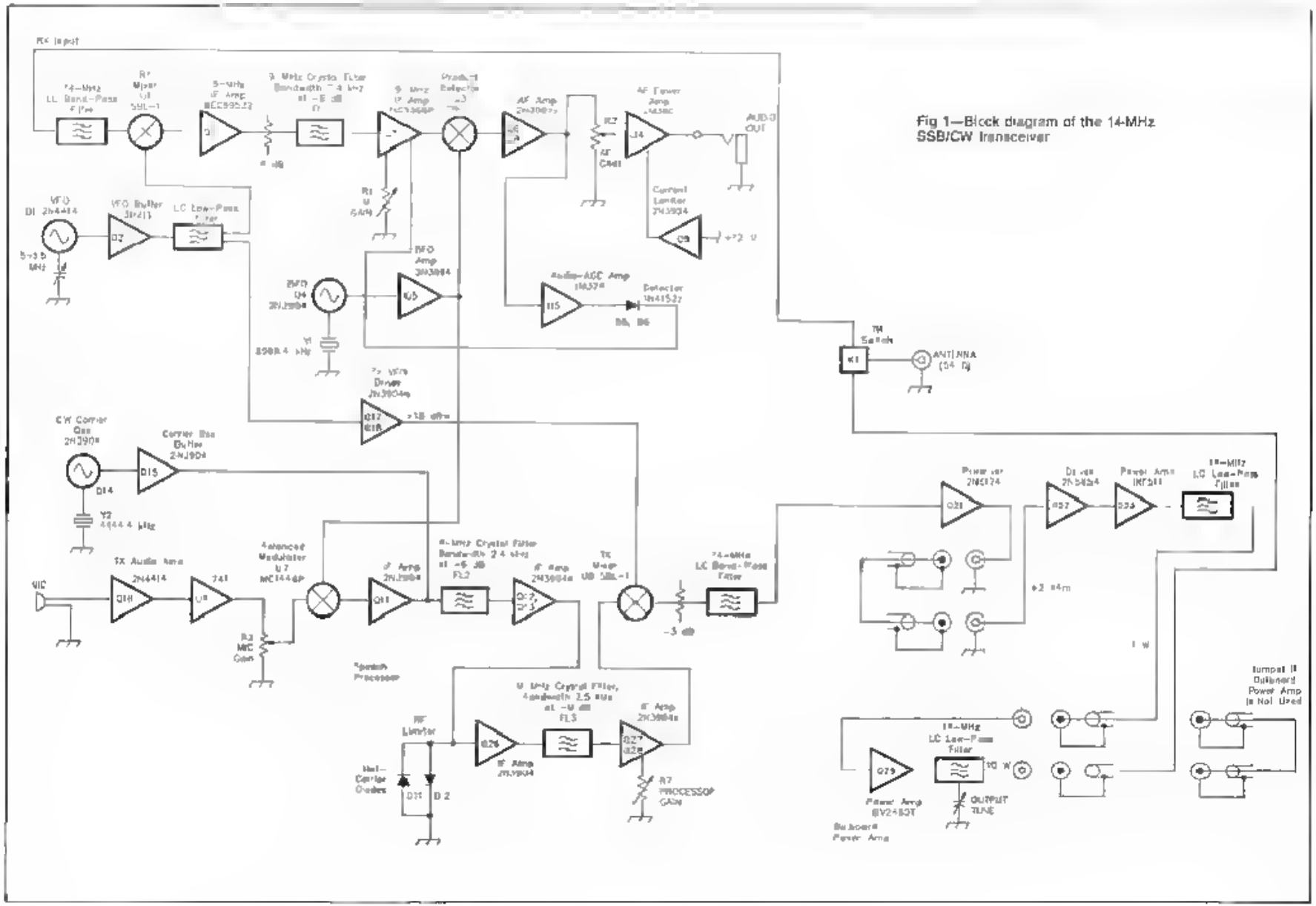


Fig 1—Block diagram of the 14-MHz SSB/CW transceiver

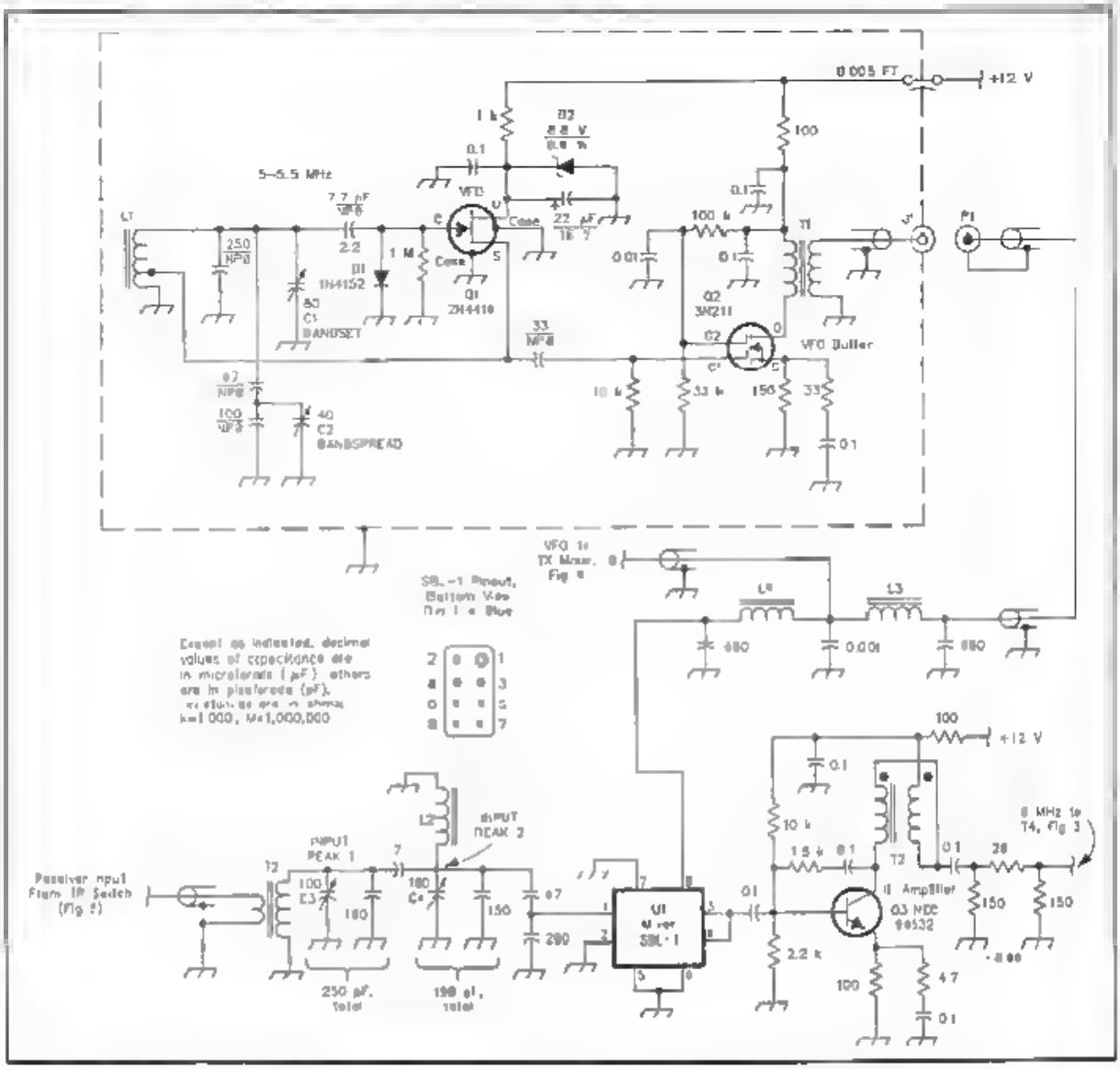


Fig 2—Schematic of the transceiver front end and VFO. Resistors are ¼ W, carbon film; unless otherwise indicated, capacitors are monolithic or disc ceramic. The VFO circuitry is built into a die-cast aluminum box.

C1, C2—Panel-mountable, air-dielectric variable with ¾-inch-diam shaft.
C3, C4—100-pF ceramic- or mica-dielectric trimmer.
J1—Coaxial jack. (The prototype transceiver uses a panel-mount SMB jack here, but a BNC or phono jack is suitable.)
L1—23 turns of no. 22 enam. wire on a T-68-6 toroidal, powdered-iron core, with a feedback tap 5 turns from the grounded end of the winding.
L2—11 trifilar turns of no. 24 enam. wire on a T-44-6 toroidal, powdered-iron core.
L3, L4—26 turns of no. 24 enam. wire on a T-37-6 toroidal, powdered-iron core.
T1—Broadband transformer. Primary, 16 turns of no. 26 enam. wire on an FT-37-43 toroidal ferrite core; secondary, 4 turns of no. 26 enam. wire wound over the primary.
T2—Narrow-band transformer. Tuned winding, 11 turns of no. 24 enam. wire on a T-44-6 toroidal, powdered-iron core. Input link, 2 turns of no. 24 enam. wire over the tuned winding's grounded end.
T3—Broadband transformer. 10 bifilar turns of no. 26 enam. wire on an FT-37-43 toroidal, ferrite core. Observe phasing.

detection by D5, USD functions as a unity gain inverter to drive a second diode (D6), providing full-wave detection. Each diode operates as a peak detector, providing one sample of the audio level per cycle. Full-wave operation doubles the sampling rate to better approach the Nyquist criterion. The practical result is a simple circuit with better dynamic performance than other audio-derived ones I've tried.

Notes

[1]W. Hayward and D. DeMaw, *Solid-State Design for the Radio Amateur* (Newington: ARRL, 1986).
[2]K. Kleinschmidt, ed, *The 1990 ARRL Handbook* (Newington, ARRL, 1989).
[3]R. and W. Hayward, "The Ugly Weekender," *QST*, Aug 1981, pp 19-21.
[4]G. Breed, "A New Breed of Receiver," *QST*, Jan 1983, pp 16-23.
[5]W. Hayward, "Designing and Building Simple Crystal Filters," *QST*, Jnl 1987, pp 24-29.
[6]W. Hayward, "A Unified Approach to the Design of Crystal Ladder Filters," *QST*, May 1982, pp 21-27; also see Feedback, *QST*, Jnl 1987, p 41.
[7]W. Hayward and J. Lawson, "A Progressive Communications Receiver," *QST*, Nov 1981. Also see Feedback, *QST*, Jan 1982, p 47; Apr 1982, p 54; and Oct 1982, p 41. This receiver also appears in the 1982 through 1990 editions of *The ARRL Handbook*.

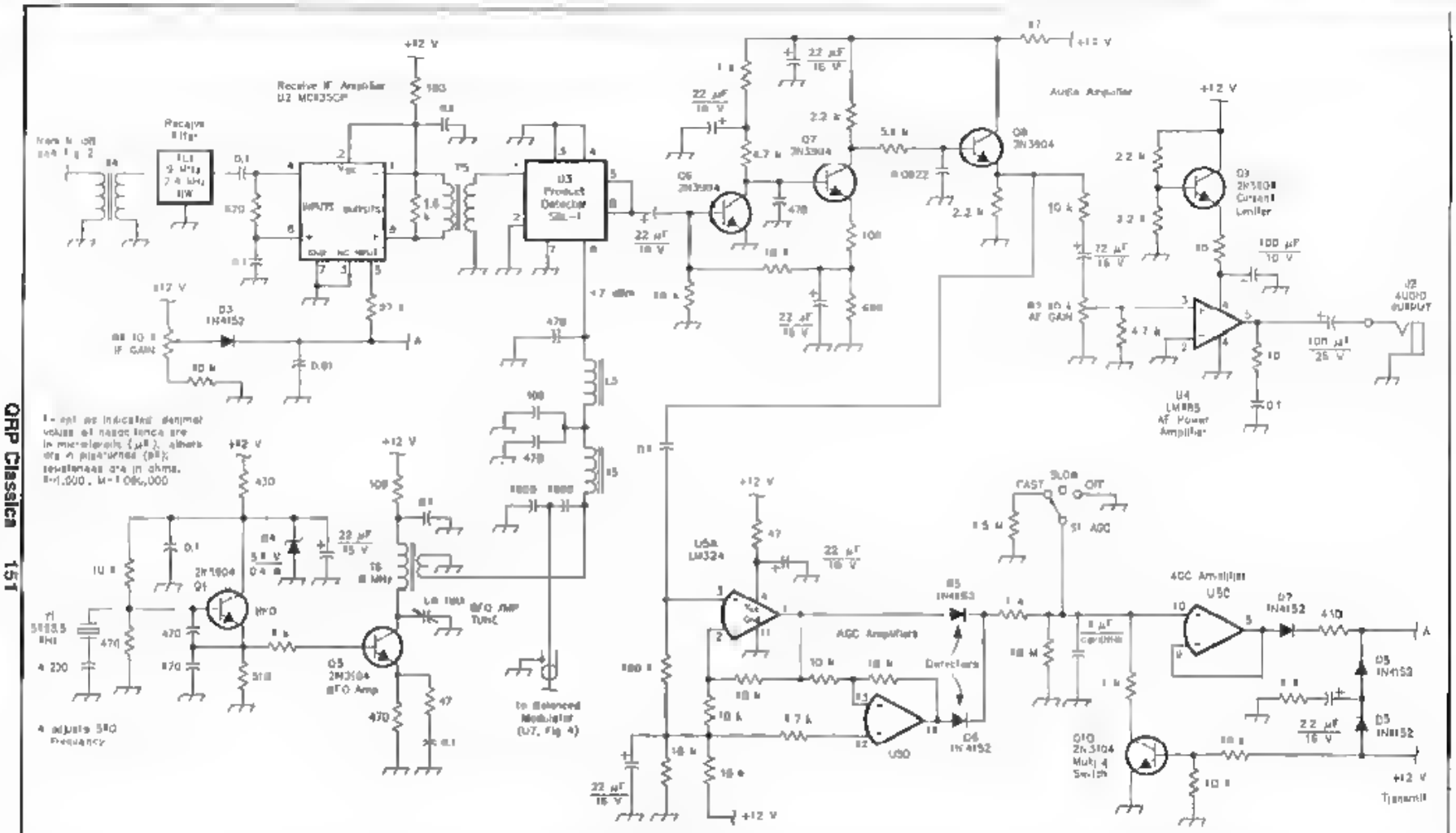


Fig 3—The transceiver receive filter, IF amplifier and detector circuitry. Resistors are ¼ W, carbon film; unless otherwise indicated, capacitors are monolithic or disc ceramic.

FL1—9-MHz crystal filter, 2.4 kHz wide at −6 dB (KVG XF-9B).

L5, L6—22 turns of no. 26 enam wire on a T-37-6 toroidal, powdered-iron core.

T4—Broadband transformer. Primary, 5 turns of no. 26 enam wire wound over the secondary winding; secondary, 16 turns of no. 26 enam wire on an FT-37-43 toroidal, ferrite core.

T5—Broadband transformer. Primary, 20 turns of no. 26 enam wire on an FT-37-43 toroidal, ferrite core; secondary, 3 turns over the primary.

T6—Narrow-band transformer. Primary, 26 turns of no. 24 enam wire on a T-50-2 toroidal, powdered-iron core.

Y1—8998.5-kHz crystal (KVG XF-901 suitable).

From January 1990 QST, p 28:

A QRP SSB/CW Transceiver for 14 MHz

Part 2: This month, W7ZOI rounds out his description of a 1- or 10-W SSB/CW rig with details on its transmitter, TR switching and optional speech processor.

By Wes Hayward, W7ZOI
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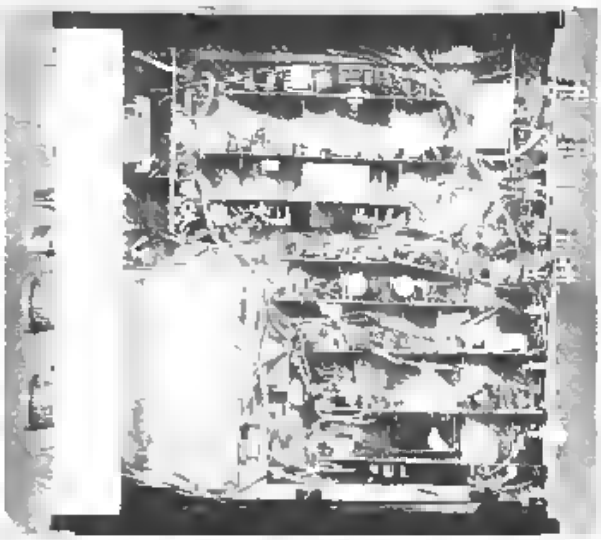
Beaverton, OR 97005



Most of the transceiver's componants are contained in ugly-constructed modules, with the exception of the VFO (lower left) and panel-mounted controls and jacks

SSB generation occurs in the circuit shown in Fig 4. A microphone amplifier (Q16-U6) supplies audio to an MC1496 balanced modulator (U7). One-microfarad capacitors (C6-C8) are used at the output of the audio amplifier and at several positions in the balanced modulator. (Originally, 10- or 22-µF units were used, but these caused the system to respond slowly during TR transitions.)

The modulator output is applied to Q11, a 2N3904 IF amplifier. This stage terminates the transmitter crystal filter and provides a convenient place for CW carrier injection. Another IF amplifier (Q12-Q13) follows the crystal filter. The TX GAIN control, R5, is set for an output of −10 dBm from Q13. This level is applied to the transmit mixer, or to the speech processor described later.

Fig 5 shows more of the transmitter. SSB energy at −10 dBm drives the transmit mixer, U8, another diode ring mixer. The 5-MHz VFO signal is amplified to +10 dBm for the mixer by Q17 and Q18. A 3-dB pad terminates the mixer, with the signal continuing to a three-pole, LC, band-pass filter. The first stage in the output amplifier chain is Q21, a 2N5179 feedback amplifier with an output of +2 dBm. This signal is looped through a coaxial-cable jumper on the transceiver rear panel for use with VHF transverters.

The driver, Q22, a 2N5859, is capable of about +20 dBm output. Transceiver output is obtained from Q23, an IRF511 HEXFET PA operating at the 1 watt output level. This power level is a little low for use on the air with dipoles, but is too high for many transverter applications.

Louder-Signal Options

Two additional circuits, shown in Fig 6, round out the SSB system. The first, at Fig 6A, is an IF speech processor. The processor is driven with a −10 dBm signal. This signal is clipped with parallel, reverse-connected, hot-carrier diodes (D11 and D12). The intermodulation products generated by the clipping are rejected by an additional crystal filter (FL3). The signal is then amplified back to the original −10 dBm level by Q27 and Q28. This circuit generates about 10 dB of clipping. Reports and measurements made on the clipped signal indicate good quality, a potential problem area with many speech processing systems.

Fig 6B shows a 10-watt-output FET power amplifier. The FET that I used (an M/A-COM DV2810) is no longer available, but is similar to the Motorola MRF138. Alternatively, one could obtain several watts of output from another IRF511.[*] The TR switching in the transceiver is set up for an onboard PA.

Summary

This was a very enjoyable project, and one that I would recommend to other experimenters. The 20-meter phone band, however, can be a little intimidating for the QRP enthusiast.

A rig like this can probably be built and adjusted by those with only modest test equipment. A 15-MHz oscilloscope served as my test equipment workhorse during construction of this project. A home-brew spectrum analyzer also served as a very useful tool, but is not required. Fig 7 shows the transceiver's CW output spectrum. The 270-Ω resistor and 0.56-µF capacitor associated with the base of Q20, Fig 6, provide CW rise and fall times of 1 and 1.5 ms, respectively. Careful measurement of signal levels during construction helps to keep the system spectrally clean.

[*] W Hayward and J Damm, "Stable HEXFET RF Power Amplifiers," Technical Correspondence, *QST*, Nov 1988, pp 36-39.

Acknowledgments

The author gratefully acknowledges the photographic assistance of Doc Lynch, KA7NPN, and technical discussions with Jeff Damm, WA7MLH.

Fig 4—The transceiver SSB generator. Resistors are ¼ W, carbon film unless otherwise indicated, capacitors are monolithic or disc ceramic.
C9—60-pF, ceramic-dielectric trimmer.
C10, C11—35-pF, ceramic-dielectric trimmer.
FL2—9 MHz transmit filter, 2.5 kHz wide at −6 dB (KVG XF-9A).
T7—Broadband transformer. Primary: 10 bifilar turns of no. 28 enam wire on FT-37-43 toroidal ferrite core, secondary, 3 turns of no. 28 enam wire over the primary. Observe phasing.
Y2—8998.5 kHz crystal (KVG XF-901 suitable).

Fig 5—The transmit mixer, driver, final amplifier and associated circuits. Unless otherwise indicated, resistors are ¼ W, carbon film, and capacitors are monolithic or disc ceramic.
C12-C14—60-pF mica- or ceramic-dielectric trimmer.
D13—1-A, 600-PIV diode.
K1—12-V dc relay.
L7-L9—20 turns of no. 24 enam wire on a T-44-6 toroidal, powdered-iron core.
L10—15-µH choke.
L11-L13—14 turns of no. 24 enam wire on a T-50-6 toroidal, powdered-iron core.
T8—Broadband transformer. Primary: 15 turns of no. 28 enam wire on an FT-37-43 ferrite, toroidal core. Secondary: 4 turns of no. 28 enam wire over the primary.
T9, T10—Broadband transformer. 10 bifilar turns of no. 28 enam wire on an FT-37-43 ferrite, toroidal core. Observe phasing.

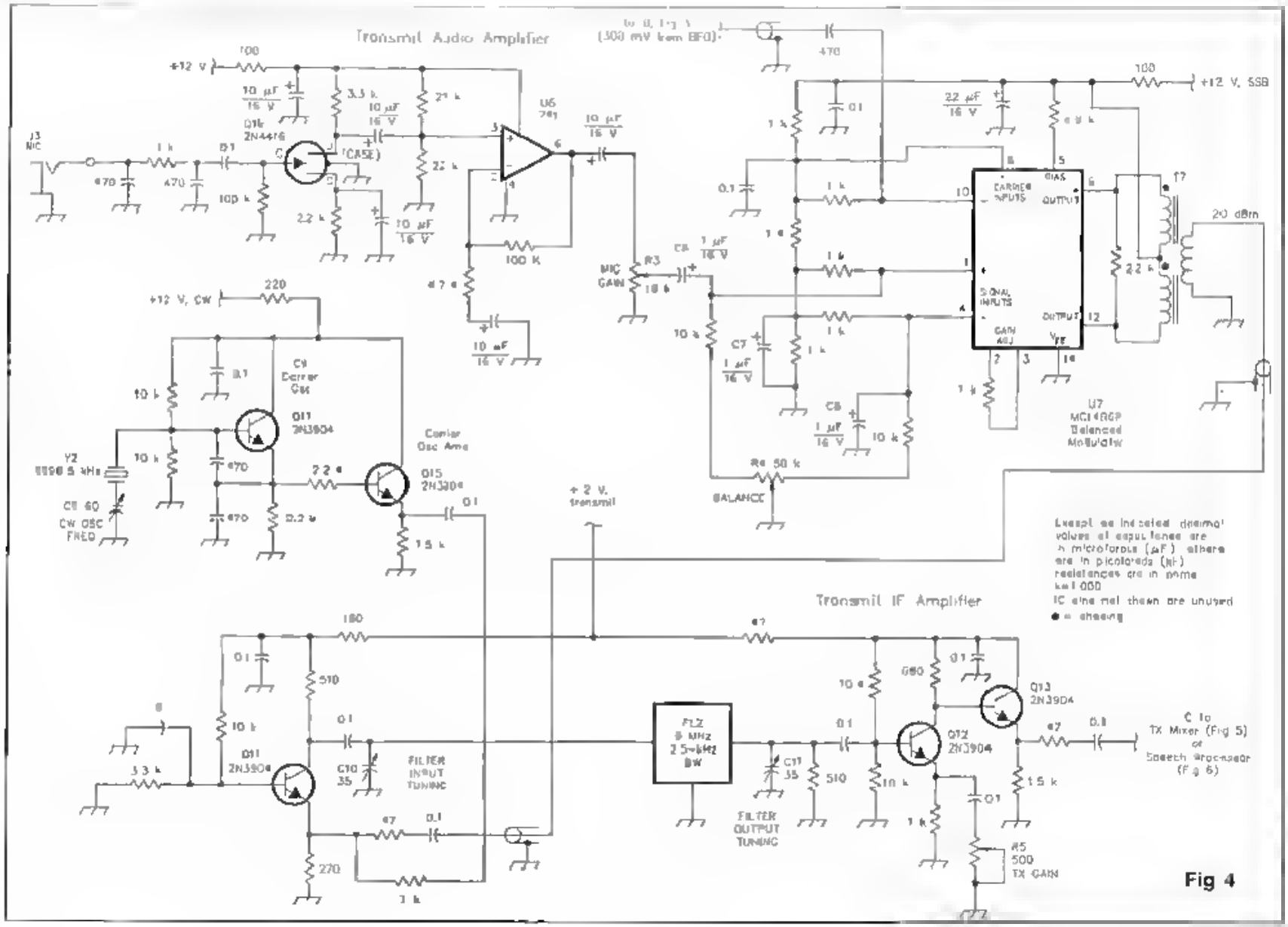

Transmit Audio Amplifier
(300 mV from BFO)
+12 V
+12 V, SSB
+12 V, CW
20 dBm
MIC GAIN
BALANCE
CARRIER INPUTS
SIGNAL INPUTS
BIAS
OUTPUT
GAIN ADJ
U7
MC1496P
Balanced
Modulator
Carrier Osc Amp
CW OSC FREQ
+ 2 V. transmit
Except as indicated decimal values of capacitance are in microfarads (µF) others are in picofarads (pF) resistances are in ohms k=1 000
IC pins not shown are unused
● = shading
Transmit IF Amplifier
FILTER INPUT TUNING
FILTER OUTPUT TUNING
9 MHz 2.5-kHz BW
R5 500 TX GAIN
C to TX Mixer (Fig 5) or Speech Processor (Fig 6)

Fig 4

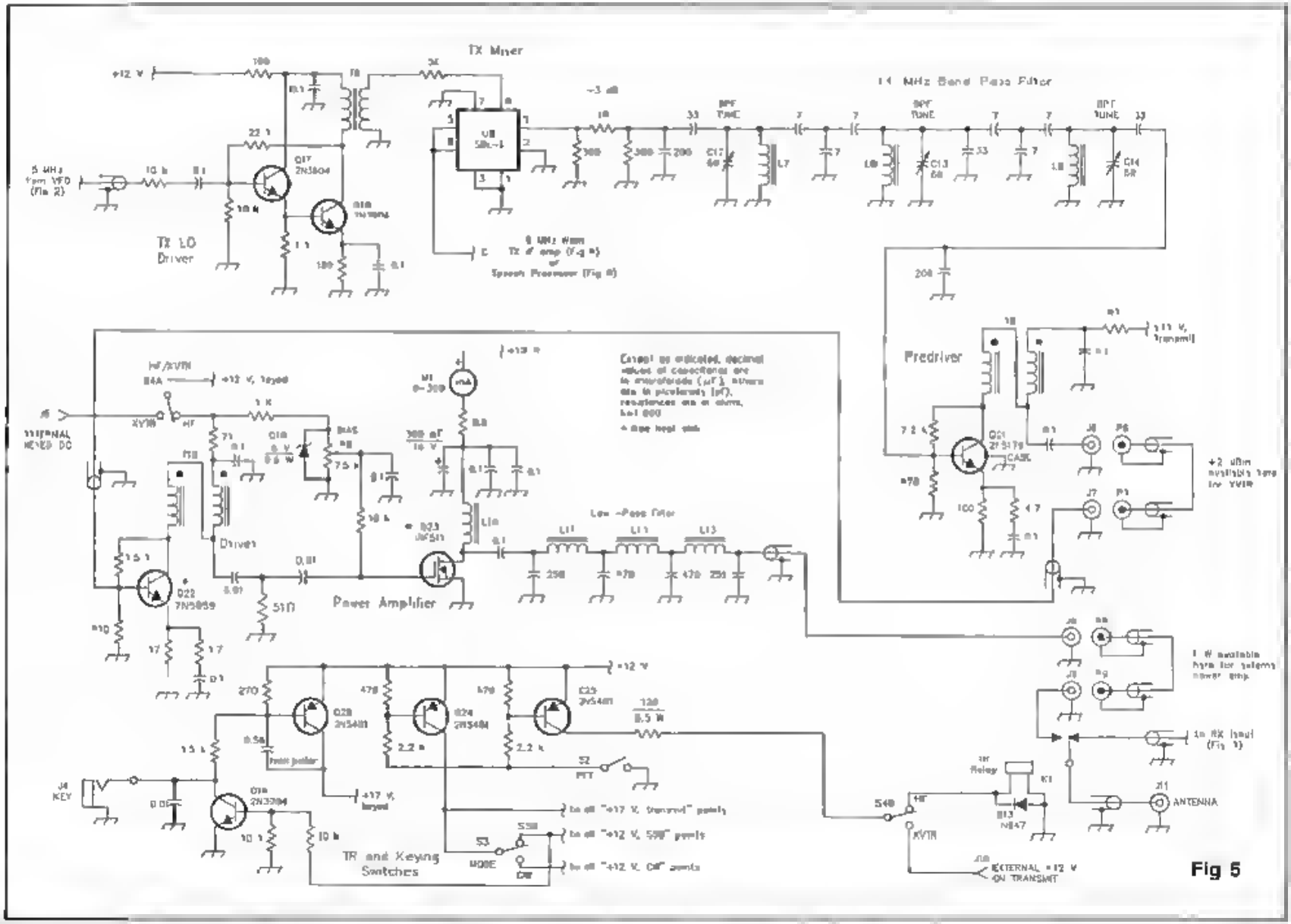
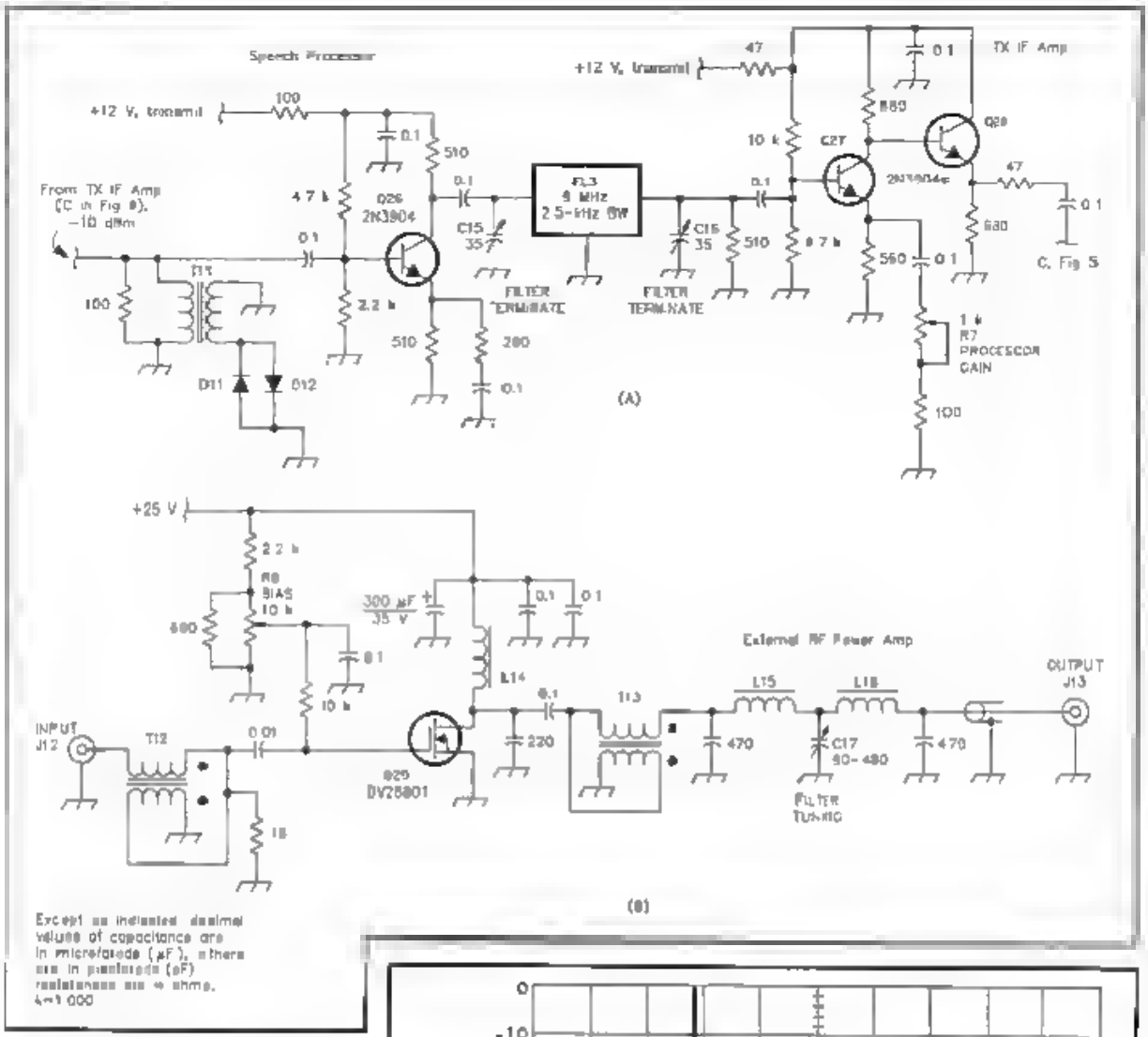

TX Mixer
14 MHz Band Pass Filter
Tx LO Driver
Predriver
Power Amplifier
Low-Pass Filter
TR and Keying Switches
ANTENNA
XVTR
EXTERNAL +12 V ON TRANSMIT

Fig 5



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF). Resistances are in ohms. k=1,000

Fig 6—The optional speech processor (A) and outboard power amplifier (B) circuits. Unless otherwise indicated, resistors are ¼ W, carbon film, and capacitors are monolithic or disc ceramic.

C15, C16—35-pF, ceramic-dielectric trimmer.

C17—90- to 480-pF, mica-dielectric trimmer.

D11, D12—Hot-carrier diode, HP-5082-2672 suitable.

FL3—9-MHz transmit filter, 2.5 kHz wide at −6 dB (KVG XF 9A).

L14—50 turns of no. 26 enam. wire on a T-68-2 toroidal, powdered-iron core.

L15, L16—19 turns of no. 20 enam. wire on a T-50-6 toroidal, powdered-iron core.

T11—Broadband transformer. Primary, 3 turns of no. 26 enam. wire over secondary; secondary, 18 turns of no. 28 enam. wire on an FT-37-43 toroidal ferrite core.

T12—Broadband transformer. 7 bifilar turns of no. 22 enam. wire on an FT-50-43 toroidal ferrite core. Observe phasing.

T13—Broadband transformer. 11 bifilar turns of no. 18 enam. wire on an FT-82-61, toroidal ferrite core. Observe phasing.

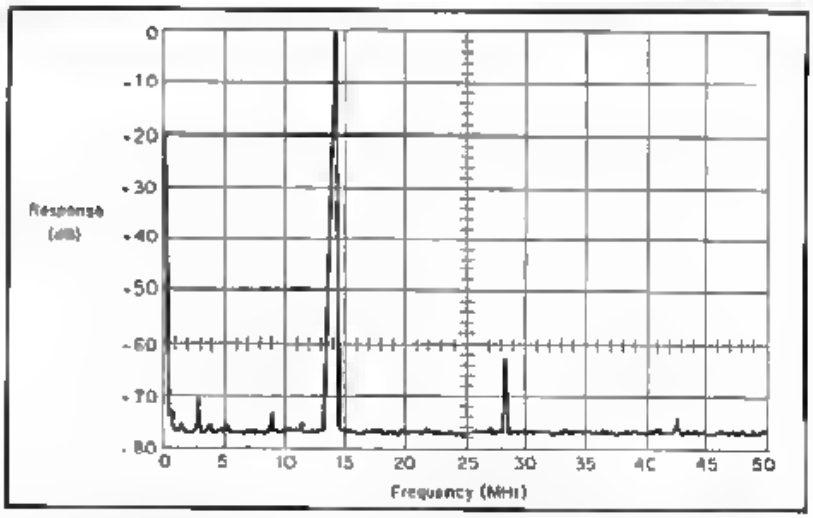
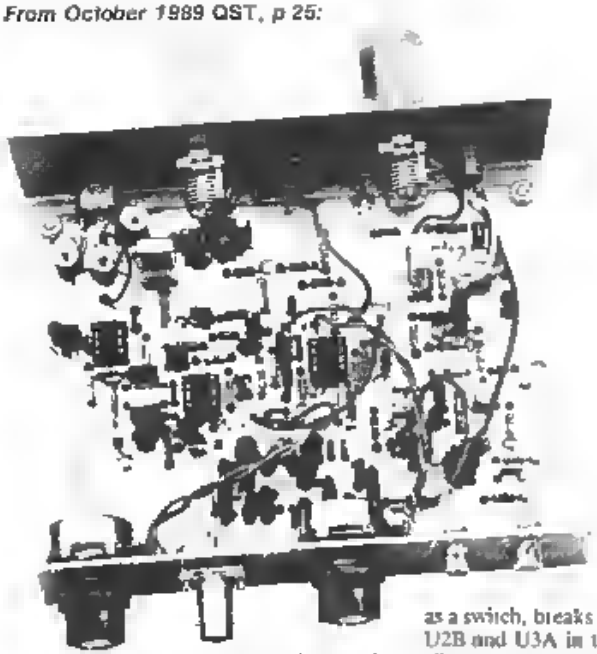


Fig 7—The transceiver's output spectrum contains a second-harmonic component 63 dB below its 1-W CW output. The major nonharmonic spurious response is a 2:1 spur near 1.5 MHz ([2 × VFO] − IF); this component is −70 dBc. The spike at far left is the spectrum analyzer's "zero-spur." An external 20-dB pad in the coaxial line provided extra protection for the analyzer, a Tektronix 2756P. The spectrum-analyzer measurements were provided by Stan Griffiths, W7NI. The transceiver complies with current FCC specifications for spectral purity.

From October 1989 QST, p 25:

The QRP Three-Bander



This low-power, direct-conversion CW transceiver covers 18, 21 and 24 MHz, and includes sidetone, spotting and relay-less full break-in—all on one circuit board!

By Zack Lau KH6CP
ARRL Laboratory Engineer

With this solar cycle's activity nearing its peak, the time for high-band QRP operation is *now*. This low-power CW transceiver is capable of exploiting these conditions. It's easy to use, sensitive enough to receive weak QRP stations, and includes audio limiting to protect your ears from loud local stations. Moving from band to band with this rig is easy: Just change crystals and re-peak its receiver input. Key down, the QRP Three-Bander produces its own sidetone—and RF, too: 1.25 to 4 watts, depending on the band, the dc supply voltage and the particular transistors used in the transmitter. And you can build the QRP Three-Bander *your* way: A complete kit of parts is available, or you can assemble your version using ground-plane construction.[1,2]

Circuit Description

Fig 1 shows the transceiver circuit. U1, an NE602N doubly balanced mixer IC, operates as a direct-conversion (D-C) product detector, converting the incoming signal directly to audio by mixing it with energy from Q2, a bipolar-junction-transistor (BJT) variable crystal oscillator (VXO). Although the NE602 achieves its conversion gain and low noise figure at the expense of dynamic range, it rejects AM broadcast-band signals well when a capacitor is present across its differential output (pins 4 and 5).

To help prevent hum pickup, the NE602's audio output is amplified by a differential amplifier (U2A, half of an NE5532 low-noise, audio-op-amp IC), which feeds a moderate-gain filter stage (U2B). The final audio-amplifier stage (U3A, half of another NE5532) drives low-impedance stereo headphones at a comfortable level. Q1, a 2N5486 junction-field-effect transistor (JFET) used as a switch, breaks the connection between U2B and U3A in transmit to keep keying clicks and thumps out of the headphones.

The QRP Three-Bander uses audio amplitude limiting instead of automatic gain control (AGC). Diodes in the filter and final-audio-amplifier stages (D1-D2, and D3-D4, respectively), and R18 (between the final audio amplifier and J2), provide ear and headphone protection by clipping the transceiver's audio output on strong signals.

Transmitter RF is generated by Q3, an MPS918 (or 2N5179) BJT operating as a VXO. Q3's output signal drives a buffer amplifier consisting of two BJTs, Q5, a 2N2222, and Q6, a 2N5109 (or selected 2N2222A). The buffer circuit is based on a design by Lewallen;[3] this version is re-biased for higher power output to make it more suitable for transmitters. The transmitter power amplifier, Q8, is an MRF237 BJT running class C. A seven-element low-pass filter (L1 through L3, and C37 through C40) reduces the harmonic content of the transmitted signal. Because this filter's cutoff frequency is high enough to pass the transceiver's 24-MHz output with little loss and yet is low enough to reduce harmonics of the rig's 18-MHz signal to a legal level, it requires no adjustment for band changes. Fig 2 shows the output spectrum of the Three-Bander's transmitter.

Full-break-in, relay-less TR switching is one of the QRP Three-Bander's finer points. The TR switch is a wide-bandwidth version of the switch used by Lewallen in his Optimized QRP Transceiver.[4] If you model this switch or measure its characteristics, you'll notice ots of passband ripple: The filter (C36, C41, C42, L4, L5 and L6) has steep skirts and three peaks corresponding to the bands covered by the transceiver. Although the calculated safe maximum-power-handling capability of this switch is just 1.4 W at 24.9 MHz, it seems to handle the transceiver output just fine. (A PIN-diode switch with appropriate biasing could handle more power, but PIN diodes are more difficult to find than ordinary switching diodes.)

The Three-Bander's transmitter section uses *differential keying*—a method of time-sequencing the keying of multiple transmitter stages to achieve a desired effect. As implemented in this circuit, differential keying helps eliminate chirp by turning on the transmit oscillator (Q3) before the buffer amplifier (Q5-Q6) comes on. This sequence is reversed at key up. The buffer amplifier turns off before the oscillator stops. Turning the oscillator on before the buffer gives the oscillator time to stabilize before the transmitter puts out RF; keeping the oscillator on after the buffer turns off assures that frequency changes by the turning-off oscillator won't be present in the transmitted signal.

To avoid key clicks—which would make the Three-Bander's signal wider than necessary for effective CW communication—the waveform of the transmitted signal is shaped in the buffer amplifier. Even though the transmitter power amplifier is nonlinear and tends to shorten the rise and fall times of its driving signal, the Three-Bander's transceiver's RF-output waveform is well-shaped, as shown in Fig 3. The open-circuit voltage at the KEY jack is positive, and about 0.5 V less than the transceiver's dc supply voltage; 1.3 mA flows in the keying circuit line when the KEY jack is shorted.

Getting the Parts

The longest part of building has nothing to do with soldering or making holes in metal. It's finding the parts! Fortunately, all the parts used in this project are sold by a number of suppliers—or you can buy a complete kit of parts from RADIOKIT, as detailed at Note 1.

Parts availability is one thing; parts cost is another. Aside from the crystals, variable capacitors C1 (FIX PEAK), C22 (FIX FREQ)

*Notes appear at end of article

The QRP Three-Bander: Vital Statistics

The performance of the QRP Three-Bander varies with band, dc supply voltage and the particular active devices used. Two versions of the Three-Bander exhibit a receive sensitivity (minimum discernible signal, or MDS) between −124 and −128 dBm, and 3rd-order IMD dynamic ranges between 71 and 74 dB. A third Three-Bander exhibits an MDS between −112 and −120 dBm, and a 3rd-order IMD dynamic range between 67 and 69 dB. Operated at 13.8 V and using an MPS918 at Q3, two QRP Three-Banders produce 2.6 and 4.0 W at 18 MHz, 2.6 and 3.4 W at 21 MHz, and 1.7 and 2.5 W at 24 MHz. A third Three-Bander (with a hand-picked 2N3904 at Q3) produces 3.6 W at 18 MHz, 3.1 W at 21 MHz and 2.4 W at 24 MHz when operating at 13.8 V. Operating the Three-Bander at 13.8 V provides 3 to 70% more transmitter output power than that available with a 13.0-V supply.

Although the Three-Bander's receiver isn't unduly sensitive or crunch-proof, it's adequate for routine amateur communication. I had no difficulty in making 3rd-order-IMD dynamic-range measurements on the Three-Bander's receiver at the ARRL lab's standard 20-kHz spacing.

The frequency swing afforded by the Three-Bander's VXOs varies with the band, stray capacitances and the particular crystals and VXO tuning capacitors used. The crystals I used allowed swings of 8.9 to 16.2 kHz at 18 MHz, 8.4 to 17.6 kHz at 21 MHz, and 14.1 to 23.4 kHz at 24 MHz.—*KH6CP*

and C28 (TX FREQ) are probably the most expensive components in this project. You can save money by purchasing these capacitors from a surplus outlet or flea market, although they are still available new. In this application, the voltage rating and physical size of C1, C22 and C28 are relatively unimportant; these capacitors need only cover the necessary capacitance range. C1 must cover the range from 15 to 45 pF. VXO capacitors C22 and C28 should have a maximum capacitance of 10 to 50 pF (10 to 15 pF is optimum) and have a minimum capacitance of just a few picofarads—the lower the minimum capacitance, the better.[3] If you can't find air-dielectric variables at an affordable price, you can replace a given variable capacitor with a switched several trimmer capacitors, as shown in Fig 1B for C1. RX PEAK; the transceiver shown in the title photo uses this arrangement. You may prefer the Fig 1B solution to C1 because flipping a switch is easier than peaking a tuning control; on the other hand, a front-panel peaking control can help you minimize interference from strong shortwave broadcasters, as discussed later in "Using the Radio on the Air." This switch-and-trimmers idea can also be applied to the transceiver VXOs; you can readjust the trimmers if your preset frequencies are occupied.

The crystal frequencies you choose depend somewhat on the particular VXO tuning capacitors you use. A VXO with a maximum tuning capacitance of many tens or even hundreds of picofarads (so much capacitance that the crystal is essentially shorted to ground with the tuning capacitance at maximum) may oscillate as much as 10 kHz below the frequency marked on the crystal. If, however, you use capacitors with maximum capacitances in the range I've specified, your VXOs should oscillate within a few kilohertz of the crystal frequency.

If you want to get your transceiver working on all three of its bands with minimal experimentation, use a 2N5109 at Q6 to ensure adequate drive to the final amplifier at 24 MHz. A *metal-cased* (TO-18) 2N2222A may work if you're willing to try several transistors at Q6 before settling on one. (I was able to use metal-cased 2N2222As in two out of the three QRP Three-Banders I've built.) If you're interested in using your Three-Bander only at 18 and 21 MHz, *any* TO-18 2N2222A will probably work at Q6.

One of home-brewing's benefits is that you can use connectors of your choice. I like to use BNC connectors as antenna jacks on HF gear. Although I don't necessarily agree with others' choices, I've seen UHF, N, and even phono connectors used for antenna connections at MF and HF. This transceiver uses phono jacks for power and keying connections. (Beware of using

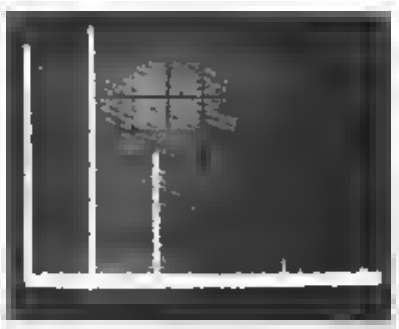


Fig 2—Worst-case spectral display of the QRP Three-Bander. Each horizontal division represents 10 MHz; each vertical division represents 10 dB. The spike at far left (the spectrum analyzer's first-local-oscillator signal) serves as a convenient "0 MHz" reference. When this spectrogram was taken, the QRP Three-Bander was producing 4 W (14.06-V dc supply) at 18.07 MHz. All harmonics and spurious emissions are at least 36 dB below peak fundamental output. The QRP Three-Bander complies with current FCC specifications for spectral purity.

Fig 1—*(Appears on following two pages.)* Schematic of the QRP Three-Bander. All resistors are ¼ W, carbon film. The inset, B, shows how to replace C1 with a switch (S2) and three trimmer capacitors (C44-C46) if you use this variation, peak C44 at 24 MHz *before* adjusting C45 and C46 for maximum received-signal strength at 21 and 18 MHz.

C1—50 pF, air dielectric, variable (Millen 21050 7- to 45-pF ceramic trimmers used in ground-plane version, and switch and trimmers [S2 and C44-C46, Fig 1B] have been used successfully). See text.
C2, C24—0.0056 μF, ceramic.
C3, C8, C10, C19-C21, C23, C27, C29, C32-C35, C43—0.022 μF, ceramic.
C4, C11, C18—1 μF, 16-V tantalum electrolytic.
C5, C6, C15—0.22 μF, metal film.
C7, C9—0.0022 μF, metal film (Bessel AF filtering). Use 0.0027 μF for tighter, Chebyshev filtering.
C12, C13—0.010 μF, metal film (Bessel AF filtering). Use 0.012 μF for tighter, Chebyshev filtering.
C14—0.056 μF, metal film.
C16—0.0015 μF, polypropylene.
C17, C31—4.7 μF, 16-V tantalum electrolytic.
C22, C28—14 pF, air dielectric, variable (Millen 21015 capacitors and Johnson 189-series trimmers have been used successfully. Several months of on-air use caused severe wear in this ex FREQ Johnson 189-series capacitor, however; use a tuning, not trimmer, capacitor at RX FREQ for longest capacitor life.) See text.
C25—0.015 μF, metal film.
C26, C30—0.082 μF, metal film.
C36, C42—15 pF, silver mica.
C37—100 pF, silver mica.
C38, C39—240 pF, silver mica.
C40—120 pF, silver mica.
C41—220 pF, silver mica.
C44-C46—30-pF plastic trimmer (optional) used in conjunction with S2, these capacitors replace C1. See text.
D1-D9—1N914 or 1N4148 silicon switching diode.
J1—Phono jack.
J2—1/8-inch stereo phone jack.
J3—BNC jack. See text.
L1, L3—Toroidal inductor, 12 turns of no. 26 enam wire on a T-25-6 powered-iron core (0.48 μH).
L2—Toroidal inductor, 13 turns of no. 26 enam wire on a T-25-6 powdered-iron core (0.52 μH).
L4, L6—Toroidal inductor, 26 turns of no. 26 enam wire on a T-44-2 powdered-iron core (3.8 μH).
L5—Toroidal inductor, 9 turns of no. 24 enam wire on a T-25-6 powdered-iron core (0.28 μH).
Q1—2N5486 JFET. A 2N4416, 2N5485, 2N5484 or MPF102 should also work.
Q2—2N3904 BJT.
Q3—BJT, MPS918 or 2N5179 recommended. A 2N3904 will work, but may exhibit faster keying rise and fall times at 24 MHz than those shown in Fig 3, and the transceiver's RF power output may be suboptimal.
Q4, Q7—2N3906 BJT.
Q5—2N2222 BJT. A 2N3904 will also work.
Q6—2N5109 RF power BJT. A selected metal-cased (TO-18) 2N2222A will work, some 2N2222As may not have enough gain for optimum power output at 24 MHz. See text.
Q8—Motorola MRF237 RF power BJT.
R1, R6, R17, R18, R21, R27, R41—100 Ω.
R2, R3, R12—2.2-kΩ (Bessel AF filtering).

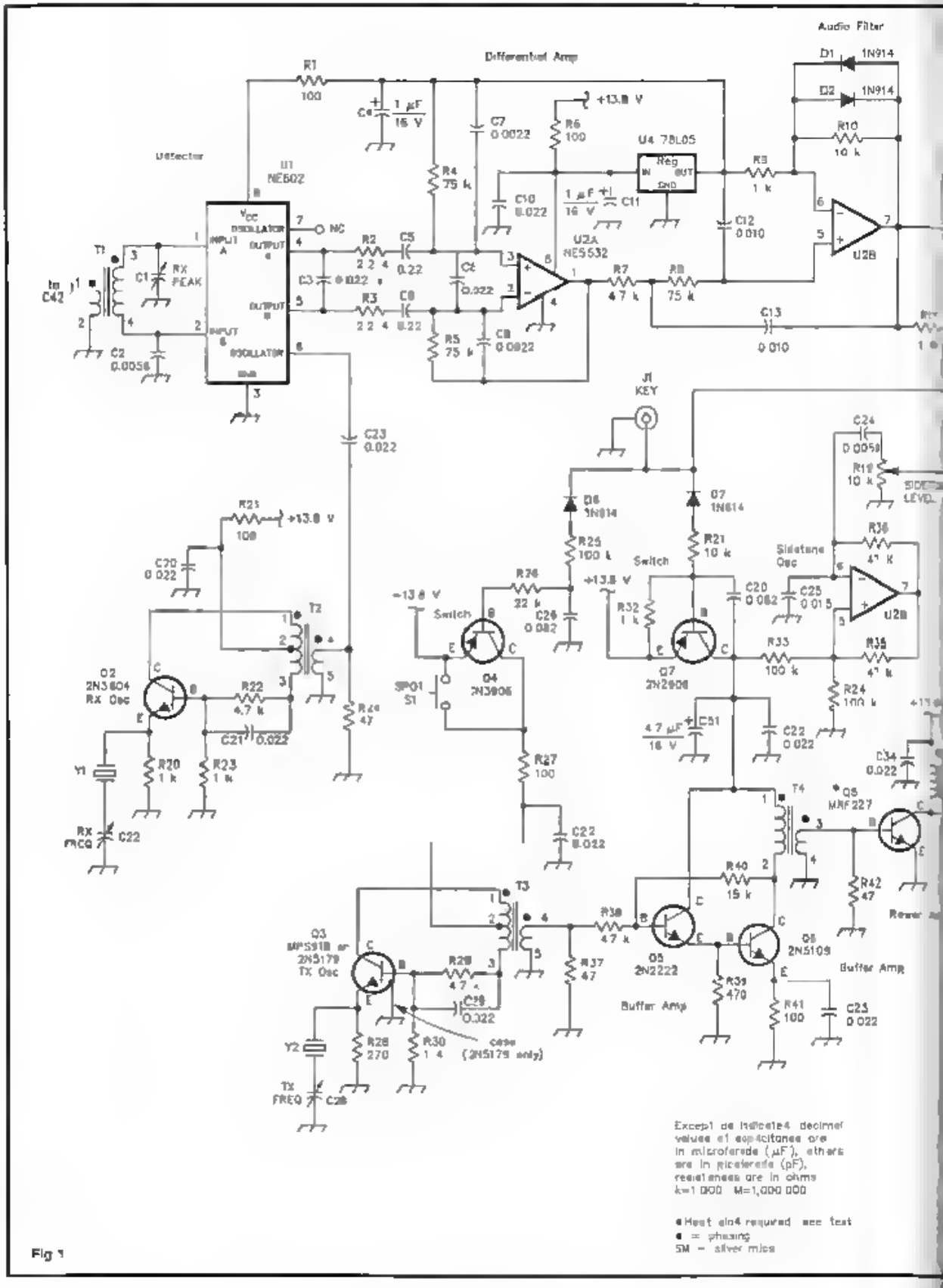
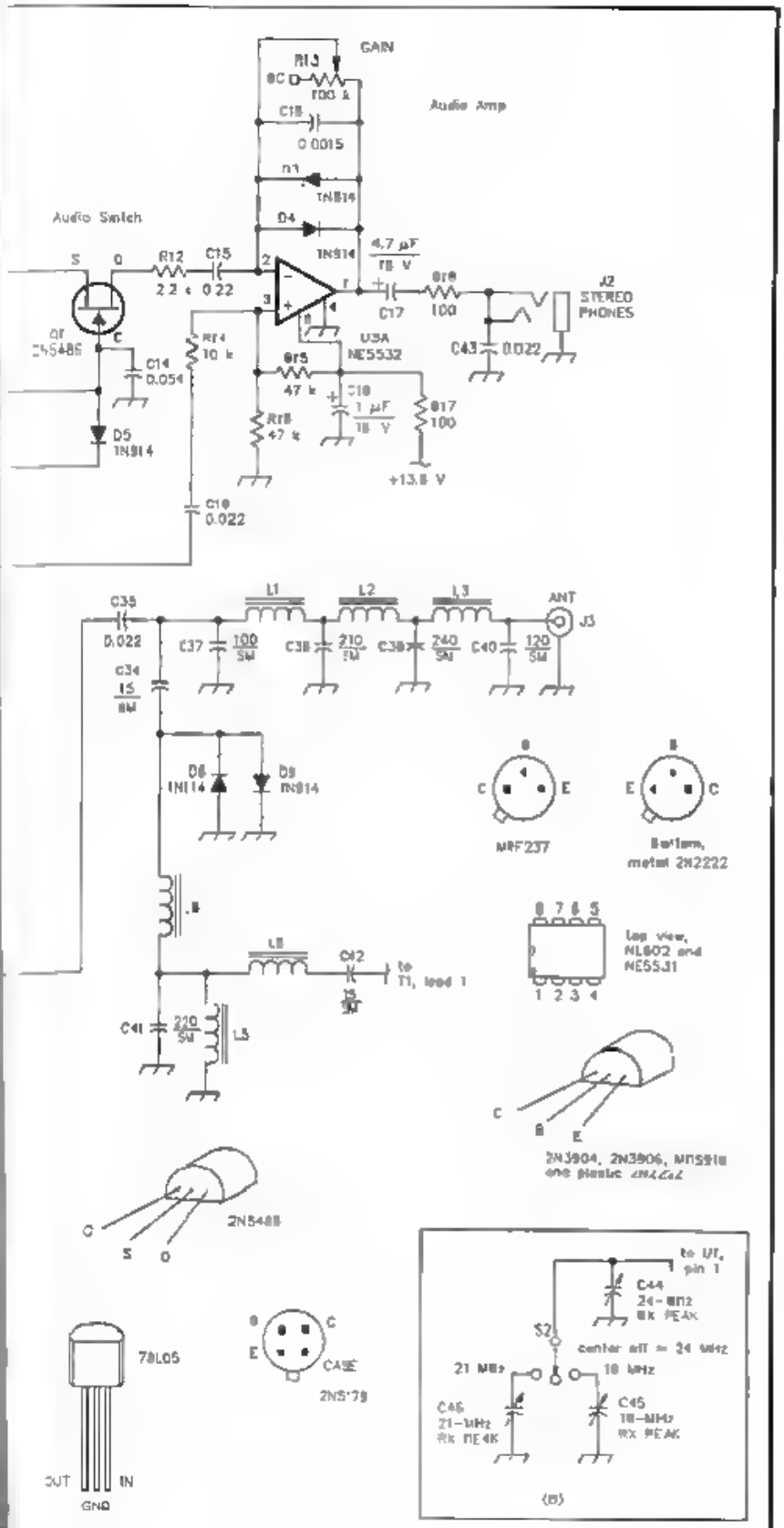

Audio Filter
Differential Amp
Detector
U1 NE602
U4 78L05
U2A NE5532
U2B
Sidetone Osc
Switch
Buffer Amp
Q2 2N3904 RX Osc
Q3 MPS918 or 2N5179 TX Osc
Q4 2N3906
Q5 2N5109
Q6 MRF227
Q7 2N2906
Q8 2N2222
T1
T2
T3
T4
J1 KEY
SPDT S1
RX FREQ
TX FREQ
Y1
Y2
D1 1N914
D2 1N914
D6 1N914
D7 1N914
+13.8 V
-13.8 V
Except as indicated, decimal values of capacitance are in microfarads (µF), others are in picofarads (pF), resistances are in ohms k=1,000 M=1,000,000
* Heat sink required, see text
• = phasing
SM = silver mica

Fig 1



Use 3.3-kΩ for tighter Chebyshev filtering

R4, R5—75 kΩ

R7, R22, R29, R38—4.7 kΩ

R8—75 kΩ (Bessel AF filtering) Use 56-kΩ for tighter, Chebyshev filtering

R9, R20, R23, R30, R32—1 kΩ

R10, R14, R31—10 kΩ

R11—1 MΩ

R13—100-kΩ, audio-taper potentiometer

R15, R16, R35, R36—47 kΩ

R19—10-kΩ, audio-taper potentiometer.

R24, R37, R42—47 Ω

R25, R33, R34—100 kΩ

R26—22 kΩ

R28—270 Ω.

R39—470 Ω

R40—15 kΩ

RFC1—Toroidal RF choke. Use 6 turns of no. 26 enam wire on an FT-37-43 ferrite toroid (10 µH)

S1—Normally open, momentary push-button.

S2—SPDT, center-off toggle (optional) Use only if C1 is replaced with C44, C45 and C46. See text

T1—Narrow-band transformer. 10:5:1 turns ratio, 21 turns of no. 26 enam wire on a T-30-6 powdered-iron toroid (primary, 1.75 µH). Secondary has 2 turns of no. 24 or 26 enam wire over primary winding

T2—Broadband transformer, 10:1 turns ratio, 20 turns of no. 26 enam wire on an FT-37-43 ferrite toroid (primary). Tap is 13 turns from the collector. Secondary has 2 turns of no. 24 or 26 enam wire over primary winding

T3—Broadband transformer, 5:1 turns ratio, 20 turns of no. 26 enam wire on an FT-37-43 ferrite toroid (primary), tap at 13 turns from the collector end. Secondary has 4 turns of no. 24 or 26 enam wire over primary winding.

T4—Broadband transformer, 3:1 turns ratio 9 turns of no. 26 enam wire on an FT-37-43 ferrite toroid (primary). Secondary has 4 turns of no. 24 or 26 enam wire over primary winding

U1—NE602 mixer IC

U2, U3—NE5532 dual low-noise op-amp IC.

U4—78L05 5-V regulator IC.

Y1, Y2—Fundamental crystal, HC-25/U holder, parallel resonance, 20- or 32-pF load capacitance. See text for discussion of frequency choice. Available from International Crystal Mfg Co, 701 W Sheridan, PO Box 26330, Oklahoma City, OK 73126-0330, tel 405-236-3741; JAN Crystals, 2341 Crystal Dr, Ft Myers, FL 33906-6017, tel 800-237-3063, and other sources.

For updated supplier addresses see ARRL Parts Supplier List in Chapter 2.

The title photo version of the QRP Three-Bander uses the switch-and-capacitors scheme shown at Fig 1B instead of C1, and Johnson air-dielectric trimmers for RX FREQ (C22) and TX FREQ (C25). Acrylic blocks machined to take set screws and the trimmers' 3/16-inch-diameter shafts serve as tuning knobs. The board is 4 × 5-3/8 inches in size.

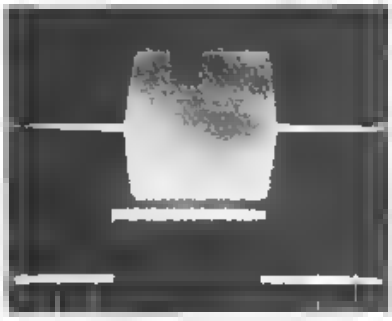


Fig 3—The QRP Three-Bander's CW-keying waveform at 10.07 MHz. The upper trace is the RF envelope; the lower trace depicts the actual key closure. Each horizontal division represents 5 ms.

phono jacks with batteries—phono plugs can short-circuit all too easily. I use Molex® connectors and fuses with my battery pigtails for safety.) The presence of identical keying and power-supply connectors isn't a problem with this rig: Nothing blows up if the key and power-supply cables are interchanged.

Construction Details

Decide early on whether you'll build the transceiver over a ground plane or on a PC board. If you decide to build a PC-board version, I highly recommend glass-epoxy (G-10 or FR-4), copper-clad circuit board over cheap phenolic board because of glass-epoxy's generally higher quality—and because glass-epoxy's greater heat tolerance allows the desoldering and replacement of components with minimal damage to the board. (This is especially important if you're new to building; you may need to fix wiring goofs.) I've made the copper pads for the wires between the board and off-board components extra large, just in case you have to do a lot of resoldering. (Small pads tend to lift off the board if subjected to too much soldering heat.) This is a trade-off in the case of the VXO-capacitor wires, though: The narrower the pads for C22 and C25 wires, the wider the VXO tuning range per crystal.

Whether you build your transceiver on a PC board or with ground-plane construction, I recommend that the transceiver circuitry be completely shielded when in use. It's important that there be grounded metal between the VXO capacitors and your fingers. Otherwise, you may experience the magic wand effect that long-time hams call *hand capacitance*. (The VXO-capacitor stators [immovable plates] are at a high impedance above ground, and anybody objects—including you—can be "seen" by those circuit points unless a grounded shield is interposed. You shouldn't be able to tune your receiver just by bringing your hand close to the tuning knob!) Complete shielding of the transceiver circuitry also helps minimize hum and microphonics in the NE602 detector, especially when the transceiver is used with a poor RF ground.

Wind your inductors and transformers before you start wiring the circuit. Amateur radio-equipment builders commonly count coil turns wrong; this usually results in coils wound with one turn too many. (Hint: With toroids, just passing the wire through the core counts as one turn.)[6,7]

Because this is an RF project, keep component leads short, as shown in the photographs. Long leads can induce excessive noise and hum into the circuit. (If you're *really* unlucky, overlong leads may cause the circuit to oscillate when it should be amplifying.) Transformers T1 through T3 aren't critical with regard to mounting—they can either lie flat or stand upright. (Some builders like to glue toroids down; I skip the glue so the coils can be removed easily if necessary.) I used screws, fork washers and ¼-inch-long metal spacers to mount my transceiver boards in their boxes.

Q8, the transmitter power amplifier, must be heat-sinked. Because the MRF237's case is connected to the transistor emitter (instead of the collector, as is usually the case with metal-cased BJTs), you can heat-sink Q8 merrily by soldering its case to the circuit-board ground foil (or to the ground plane, if you're undertaking ground-plane construction.) That's what I did in my Three-Bander. One small solder joint does the job; you needn't solder the entire case perimeter.

I used three-hole mount phono jacks (two holes for mounting screws and one for the jack barrel) because they don't loosen with use as easily as single-hole mount types do. For the same reason, I took the time to drill the extra holes necessary to seat the anti-rotation tabs on the GAIN and SIDETONE LEVEL controls because seating the tabs—instead of breaking them off—results in controls that almost never work loose from the panel!

Testing

None of the Three-Bander's circuits need be trimmed or aligned beyond the adjustments possible with its panel controls, so you need only verify that it works. I suggest powering the transceiver with a small supply during testing—a 12- to 13-V regulated power supply capable of producing no more than 0.5 to 1 A is fine. (A supply capable of some 5 or 10 A invites the possibility of serious smoke and component destruction if you make a wiring mistake. Don't use batteries, either: Short-circuited, they can source enough current to melt wires!)

The first test is to determine whether or not the Three-Bander can hear its own transmit oscillator. Set the GAIN control to the middle of its rotation. Plug in crystals

On the Air with the QRP Three-Bander

What can you expect of the QRP Three-Bander? In three brief operating periods, I snagged OK3CQR, EA8AB, OK2KFM, KK6H, W5BYBT, AF4S and W9MNU at 18 MHz, and KF5OL and G3FGT at 21 MHz—nine contacts, four countries, four states and three continents. The antenna? Fifty or so feet of wire teased in a tree and worked against a baseboard-heater "ground."

The QRP Three-Bander's receiver is more than satisfactory, considering its simplicity. There's audio to spare; I didn't have to turn the GAIN control wide open all the time. The receiver is a bit microphonic, but not annoyingly so. I heard a bit of hum at some settings of the RX PEAK control—probably because I used an un-operated power supply in conjunction with my crummy RF ground. Sometimes, I had to use RX PEAK to minimize AM "breakthrough" from strong 17- and 21-MHz broadcasters. All this means is that I'll build my version of the QRP Three-Bander with a front-panel-peakable front end. (I didn't hear one bit of breakthrough from local medium-wave broadcasters, by the way.)

Full break-in is fun with the QRP Three-Bander. Zack Lau has solved several problems at once by incorporating audio limiting into this transceiver: "De-thumping" the rig's TR switching, protecting the operator's ears and headphones from overdrive, and ridding the rig's sidetone of monotony. (In transmit, you hear the sidetone [assuming that you've set its SIDETONE LEVEL control to skew this, of course], a tone corresponding to the frequency difference between the receive and transmit VXOs, *and* the products of intermodulation between these signals as they mix in the Three Bander's audio-limiting circuitry. Result: The rig's "sidetone" rarely sounds the same two QSOs in a row!)

The QRP Three-Bander's differential keying is a class act. Listened to with my NRD-525 receiver, the Three-Bander's CW sounds absolutely A1 (pun intended) at 18, 21 and 24 MHz. (No "*Sure* the keying's too hard—but heck, it's QRP" excuses are necessary for *this* low-power rig.) If you must key an oscillator for CW, this is how to do it.

Working all continents will be easy with this rig. Who'll be first to work all *states* with a Three-Bander? It probably won't be me—at least, not unless I build mine soon: Other HQ staffers are lining up for their shots with KH6CP's QRP Three-Bander!—*David Newkirk, AK7M*

How About Modifying the QRP Three-Bander?

I'm sure that many of you would like this transceiver to cover different bands—14 or 28 MHz, for instance. The problem is that if I'd taken the time to work out the details of all such permutations before publishing this article, you'd never have seen this article! The first step in getting a successful equipment design off the drawing board and into reality is defining the limits of what you want to accomplish—so I decided to design the QRP Three-Bander to cover only the 18, 21 and 24-MHz amateur bands. That said, though, I do have some untested Three-Bander-modification hints for the adventurous.

First of all, most of the QRP Three-Bander's RF circuits are broadbanded enough to cover the HF spectrum without modification. The exceptions are the NE602's tuned input circuit (C1-T1), the transmitter-output low-pass filter (C37-C40, L1-L3), and the TR switch filter (C36, C41, C42, and L4-L6). The variable crystal oscillators should work fine from 3.5 to 28 MHz with fundamental-mode crystals, keep in mind, however, that a VXO's tuning range generally *decreases* as the crystal frequency is lowered.

You may need to add a few more turns to T4's primary winding on the low bands, but T4 is pretty broadband, too.

The TR switch is a bit tricky to design—you can't get the peaks in a high-ripple band-pass filter to fall exactly where you want them merely by poking at a calculator—but you can always use Lewallen's single-band version (see Note 4 of the main text). If you use a rotary switch for the TR circuit, receiver front end and transmitter output filter, getting the QRP Three-Bander to cover five or six bands shouldn't be too difficult. —*KH6CP*

at Y1 and Y2, both on the same frequency, and set the **RX PEAK**, **RX FREQ** and **TX FREQ** capacitors so that their plates are half meshed. Press the **SPOT** button and adjust **RX FREQ** to find the transmitter signal. If you're lucky, you now have an expensive code-practice oscillator! If you can't hear anything at all, even after trying the entire range of the **GAIN** control, there's a fault in the receiver audio chain.

If you can meter your power supply's output current, you can confirm that the transmitter works by attaching a 50-Ω dummy load to (a 2-W carbon or metal-oxide resistor) and briefly shorting the **KEY** jack. If the transmitter is working, the transceiver's current drain should increase to 200 or 300 mA with keying. You can also use an oscilloscope, RF wattmeter or RF voltmeter to measure the transmitter output.

Using the Radio on the Air

The ultimate test of a home-brew rig is making contacts. Although D-C receivers are notorious for fooling their operators into transmitting on the wrong frequency, the QRP Three-Bander's independent transmitter and receiver oscillators can help you avoid this problem. Perhaps the easiest way is to let other stations thaw you in properly by calling CQ on a clear frequency! The second way to be sure you've spotted your transmitter on the other station's frequency is to adjust **TX FREQ** so that your spotting signal (1) has the same pitch as the incoming signal and (2) "tones the same way" as the incoming signal. In other words, if, once you've spotted your transmitter, adjusting **RX FREQ** causes your spotting signal to rise in pitch as the incoming station falls in pitch, or vice versa, you've set your transmitter to the wrong frequency and must readjust it to "the other side of zero beat" to put your signal in the other station's receiver.[9] Another approach is to adjust **RX FREQ** for zero beat with the incoming signal, press **SPOT**, and adjust **TX FREQ** to zero beat your transmitter to your receiver *and* the incoming signal. Once you've done this, adjust **RX FREQ** to receive the incoming signal at the pitch you want. This latter approach works well on a busy band: After you've spotted your transmitter, readjust **RX FREQ** to receive the less-interfered-with "side" of the received signal. (Incidentally, you can zero-beat your transmitter to fairly strong incoming stations by pulling out the receive crystal [Y1] and adjusting the spotting signal to zero beat. In this case, the spotting signal acts as the receiver local oscillator. The spotting and incoming signals will have exactly the same pitch when you plug the receive crystal back in.)

Adjust **RX PEAK** for maximum received-signal strength. If no man-made signals are audible, adjust **RX PEAK** for maximum background noise. You can also use this control as an attenuator by misusing it—a useful feature when strong signals overload the rig's mixer. Usually, detuning **RX PEAK** just enough to reduce the culprit signal below the overload point preserves sufficient desired-signal sensitivity for you to keep operating.

Adjust the **GAIN** control for a comfortable listening level; this control does not affect the sidetone level. For finding stations, I adjust **GAIN** so I can just hear the background noise.

The **SIDETONE LEVEL** control adjusts what its name implies. Although a sidetone is unnecessary with a straight key, it's quite handy—if not essential—with electronic keyers.

The **SPOT** button turns on the QRP Three-Bander's transmit oscillator, allowing you to adjust your transmit frequency without actually transmitting a signal on the air. (Sweeping a signal across a band is considered poor amateur practice—even for antenna testing.)

Summary

The QRP Three-Bander gets you going on three of our best high bands in style, and with enough power to work the world. Build it, use it—and have fun!

Notes

[1] Kits of parts for the QRP Three-Bander are available from RADIOKIT, PO Box 973, Pelham, NH 03076; tel 603-437-2722; for $99 each, plus $4 each for shipping via the United Parcel Service in the US. (Canadian and overseas orders are welcome; contact RADIOKIT for details.) The kit price includes a PC board and pre-punched Tan-Tec enclosure and all QRP Three-Bander components *except crystals*. The ARRL and *QST* in no way warrant this offer.

A PC-board template and parts outline for the QRP Three-Bander are available for a business-size SASE from the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111.

[2] This technique, also known by the unfortunately pejorative term *ugly construction*, entails fastening circuit components—connected directly to each other by short leads—above a thin copper sheet (ground plane). Despite their appearance, circuits built in this way generally work better than their PC-board-built counterparts because all are better duplicates than fiberglass or phenolic. Builders well-versed in ground-plane construction can generally build the ground-plane version of a given circuit faster than its circuit-board equivalent.

[3] R. Lewallen, "An Optimized QRP Transceiver," **Feedback**, *QST*, Nov 1980, p 53.

[4] R. Lewallen, "An Optimized QRP Transceiver," *QST*, Aug 1980, pp 14-18. I highly recommend this article to anyone who wants to build a 40-meter QRP transceiver.

[5] Most of the frequency variation provided by C22 and C23 occurs at the low-capacitance end of their capacitance spans. Thus, achieving the smallest possible minimum capacitance at C22 and C23 is especially important to builders who duplicate this project will ground-plane construction because of the generally lower stray capacitances this construction method affords. Even tiny capacitances in the VXO circuit maximizes the VXO tuning capacitor's contribution to capacitance change in the circuit.

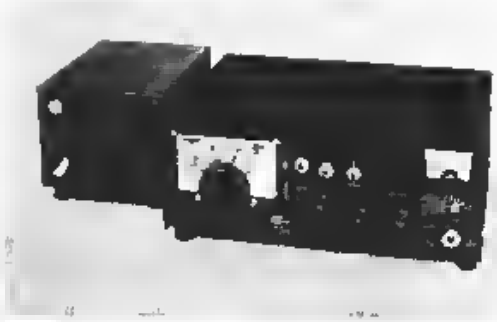
[6] Fig 7-3 on page 2-37 of the 1989 *ARRL Handbook* shows several aspects of toroid construction, including how to count toroid turns accurately, and how to wind a toroidal transformer (like T1, T2 and T3 in this project). —*Ed*

[7] The inductances listed for L1 through L6 are *measured* values. If you attempt to verify these inductances by using well-known toroid-inductance formulas to work backwards from the core and turns values given, you'll come up with different inductance values. This is so because simple formulas for calculating the inductance of toroids tend to overestimate the inductance. (Such formulas are easy to spot. They fail to take wire thickness into account, resulting the same inductance whether you use wire so thick you can barely wind it, or hairlike wire that's nearly invisible!)

[8] These are important considerations in portable QRP operation because you're usually left the necessary nighttuning tools at home!

[9] Because the days when ham radio amateurs routinely tuned for ranges over a significant portion of a band are long gone, accurately spotting your transmitter is important. Spotted on "the wrong side of zero beat" but at the same pitch as the incoming signal, your transmitter is twice that pitch away from the incoming signal—for example, 1.4 kHz away from an incoming signal tuned to produce a 700-Hz pitch. —*Ed*

From February 1975 QST, p 16



QRP TRANSCEIVER FOR 50 MHZ

Three watts PEP will do it on 6!

Part 1

BY PETER J. BERTINI,* K1ZJH

The "World Above 50 Mc" has lured an increasing number of amateurs over more than four decades. One of its main attractions has been the effectiveness of low-power gear, especially the portable variety. But with the advent of single-sideband, some of the fascination with vhf operating has tended to fade, especially on the 50-MHz scene, where 200-watt transceivers and the inevitable 2-kW linears tend to make the band seem little different from lower amateur frequencies. After all, how many stations can operate on 50.110, all at once, and not have a reminiscent of DX pileups on 20? Fortunately there is a difference – there is plenty of room for everyone in the 50-MHz band, and if we just move up a little in frequency, low power works extremely well. Here's an attractive way to get into the QRP game on 6.

AFTER GOING the high-power route for several years, this author decided that QRO was not his cup of tea and 6-meter activity fell into a rut of dreary ragchews until fairly recently. After reading numerous articles by W1CER and other exponents of the QRP cult, we could see no reason why this philosophy should be applied to the hf bands only, especially when the wide-open spaces in the vhf realm are considered. The 50-MHz band should be ideal for QRP work, as we use ssb. It offers more opportunities for DX, and a greater variety of propagation media than higher bands, and there is more general use of ssb, a mode well-adapted to QRP techniques, now that solid-state gear is "the way to go."

We set a work and the result is a small package ideally suited to Field-Day type excursions and local mobiling. It is also a very practical primary or back-up rig for the 50-MHz home station. Its robust 3-watt PEP signal is hardly distinguishable from that of stations running 500 watts or more when a good antenna system is used – and "neighbor trouble," the bane of so many urban 6-meter operators, is virtually nonexistent.

As can be seen from the first photograph, the transceiver is self-contained except for the power supply. This could have been built-in, but a separate power source is advantageous. It allows use of an ac supply at home, direct connection of a car battery for mobile work or operation from any of several types of portable batteries, including the 15-volt rechargable units often used with solid-state TV sets.

Eleven separate printed-circuit subassemblies are used, each a basic component of the transceiver circuitry. This allowed individual experimentation and trouble-shooting, and it leaves the way open for changes or additions without complete rebuilding. Board templates and layout are not critical. Construction of such a rig is beyond the capabilities of the novice builder. Experienced constructors interested in duplicating some of the items described, should have little trouble working from the individual schematic diagrams and text.

As will be seen from the block diagram, Fig. 1, the principles involved are not unlike those of crystal-filter ssb transceivers for lower frequencies but applying these principles is simplified in a one-band design. That good things come in small packages" is amply proven by the simplification resulting from use of integrated-circuit packages wherever possible. They make a marked reduction

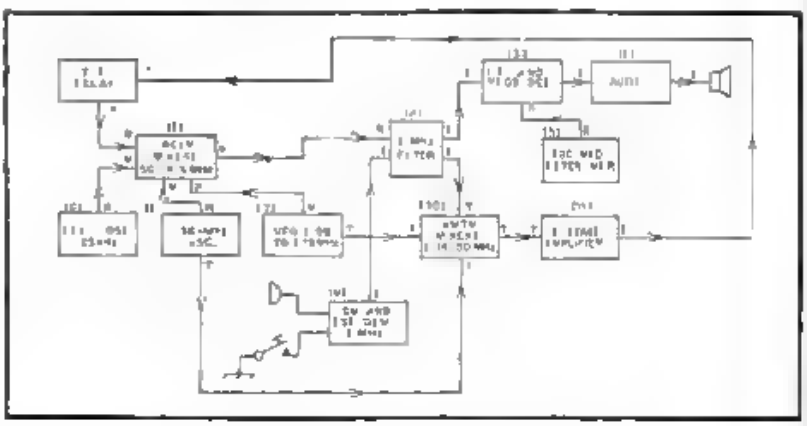


Fig. 1 – Block diagram of the K1ZJH 50-MHz transceiver. Each item shown is a separate printed-board assembly. Numbers in parentheses are the order of their description in the text. Transmitting and receiving functions are given adjacent to the signal path lines. Items 9 through 11 are described in Part 2.

* 20 Patan Road, Sussex, CT 06071

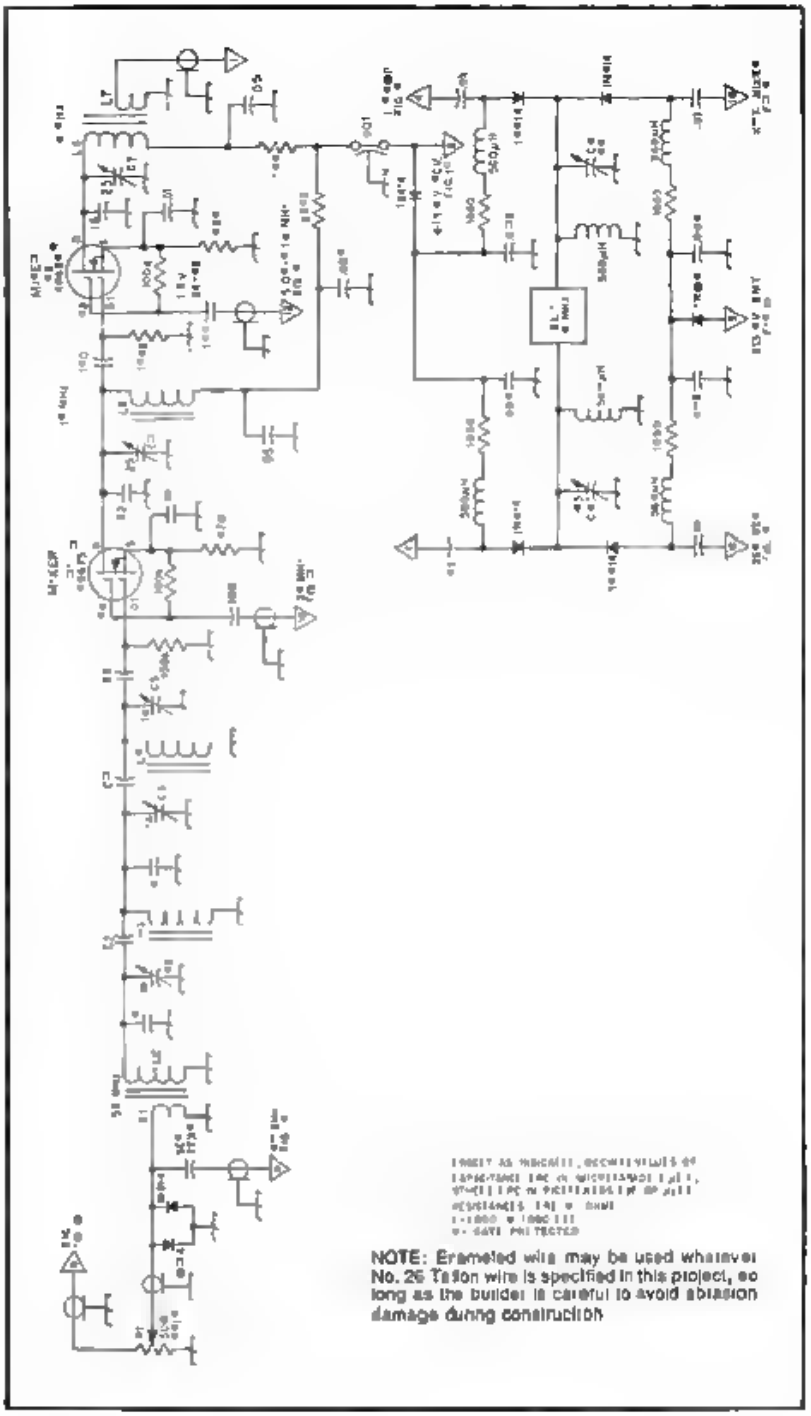
a number of parts required, in several key areas. A small crystal filter is the heart of the system, mainly responsible for the excellent performance, both transmitting and receiving. The dual conversion lineup will be discussed in step-by-step order to avoid confusion, following the numerical order indicated in Fig. 1.

Receiver Front End

Eliminating the rf amplifier stage in a "good" HF receiver is usually considered a bit unorthodox, and is frequently representative with poor performance and poor tuning. This holdover from vacuum-tube limitations is now something of an "old wives' tale," as the noise figure of a well-designed transistor mixer for 50 MHz can be lower than the external noise encountered in most amateur operation. Especially when simplicity is a factor, eliminating the rf amplifier offers a desirable trade-off between sensitivity and good strong-signal overload capability. It is interesting to note that some of the finest commercial amateur receivers for this service do not use rf amplification ahead of the mixer stage. Especially for use with a very low-powered transmitter, the ultimate in low noise figure is certainly not important in a 50-MHz receiver.

Some types of rf gain-control systems introduce nonlinearities in the receiver front end, increasing susceptibility to overloading and cross-talk. The problem is avoided with the use of a miniature 500-ohm control, R1, across the receiver input, which serves as a simple yet effective rf gain control. Though the mixer transistors, Q1 and Q2, are dual-gate MOSFETs with built-in transient-suppression diodes, additional protection is provided with IN914 diodes, connected in opposite polarity across the mixer input.

The incoming 50-MHz signal passes through three lightly coupled toroidal *LC* circuits in a simple band-pass network, for reasonable rejection of out-of-band signals. The first of two mixers converts the 50-MHz signal to 14 MHz. Injection at 36 MHz is generated by an overtone crystal oscillator used in the transmitting section for spot reception, 14 to 50 MHz. The 14-MHz IF passes through a simple *LC* network to preserve bandwidth. The signal is then mixed with the 5-MHz VFO output in a second 40673 stage, producing the second IF, 9 MHz. The injection level at gate 2 of both mixers is 1.5 V pk-pk.



NOTE: Enameled wire may be used wherever No. 26 Teflon wire is specified in this project, so long as the builder is careful to avoid abrasion damage during construction.

Top view of the 50-MHz transceiver. Circuit-board assemblies identifiable in this picture are listed by number given in Fig. 1. (1) Receiver mixers, upper left corner. (4) Audio amplifier, square assembly, upper right. (6) Calibration oscillator, left center. (8) 36-MHz oscillator, small board, lower center. (10) Transmitting mixer, long narrow assembly, right center. (11) Transmitter amplifiers, far right. The agc and meter amplifier assembly (5) is vertically mounted in the back of the case, lower left corner, in it is not clearly distinguishable at left. The VFO (7) is in an aluminum enclosure directly under the calibration oscillator, except for its tuning and band-setting capacitors, which are visible at the lower left.

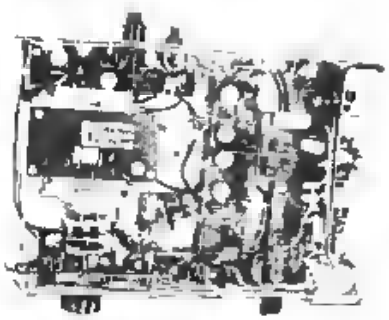


Fig. 2 — Schematic diagram and parts information for the 50-MHz transceiver front end and 9-MHz filter. Parts not described below are numbered for text reference.

C1, C3, C5 — Subminiature air trimmer, 1-14 pF.
C2, C4 — Gimmick capacitor, 2 twists No. 26 enameled wire, 1/2-inch long; approx. 0.5 pF.
C6, C7 — Ceramic trimmer, 5-25 pF.
C8, C9 — Ceramic trimmer, 7-45 pF.
FL1 — 9-MHz crystal filter (Spectrum International, Box 87, Topsfield, MA 01983, Type XF9B).
L1 — 1 turn No. 26 enameled wire on 0.37-inch toroid core (Amidon T-37, yellow).
L2 — 13 turns like L1, on same core. See text for 25-kHz coupling method.
L3, L4 — 12 turns on core like L1, L2.
L5 — 36 turns No. 32 Teflon on 0.5-inch toroid core (Amidon T-50, red).
L6 — 55 turns like L5.
L7 — 11 turns at low-Z end of L6.
Q1, Q2 — Gate protected MOSFET (RCA 40673).

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

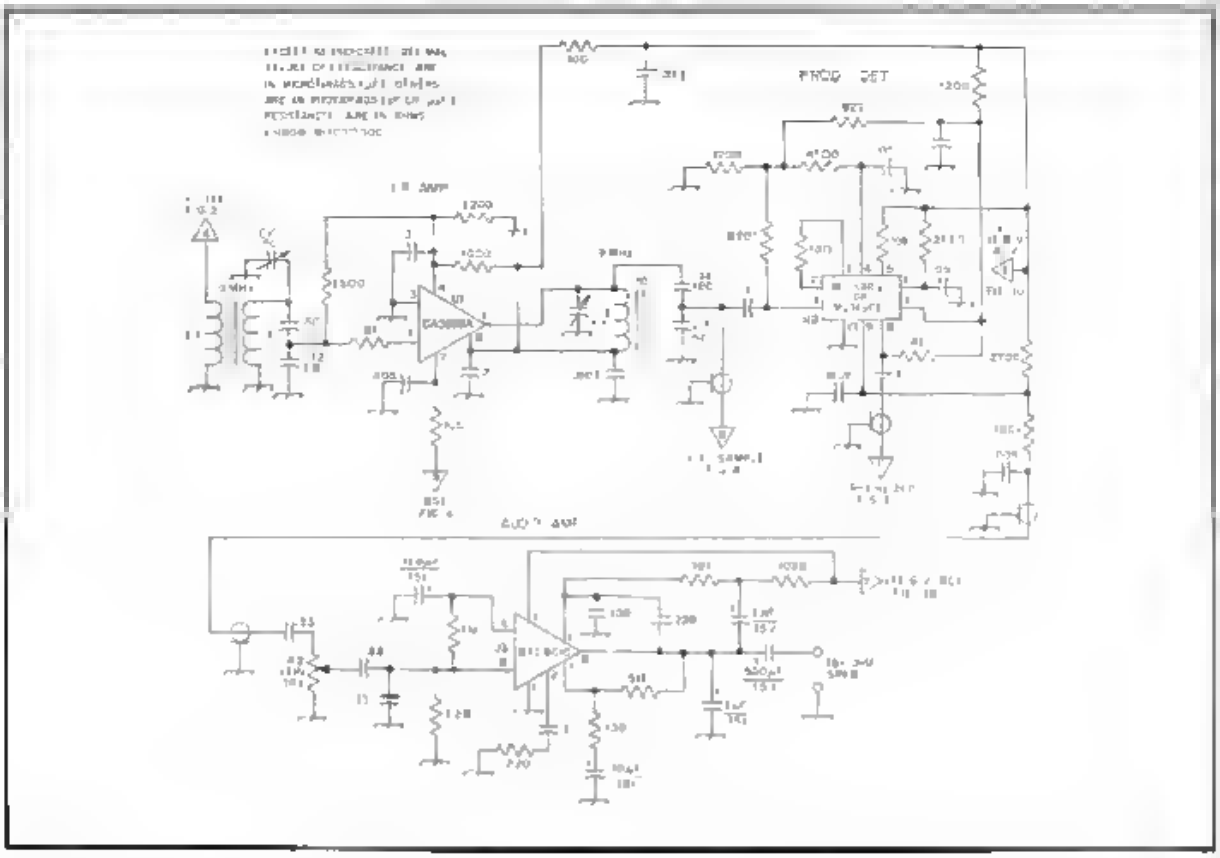


Fig. 3 — Schematic diagram of the 9-MHz i-f amplifier and product detector, and audio-amplifier modules. Parts not described are numbered for text reference.

C10, C13 — Ceramic trimmer, 5-50 pF.
L8, L9 — 17 and 24 turns, resp., No. 26 enameled wire on 0.5-inch toroid core (Amidon T-50-2).
L10 — 24 turns, like L1.
R2 — 10,000-ohm miniature control.
U1 — I-f amplifier IC (RCA CA3028A).
U2 — Product Detector IC (Motorola MC1496G or 1596G).
U3 — Audio amplifier IC (Motorola MFC8010 or µEP C8004).

considered to be optimum for conversion efficiency and mixer linearity. Output from the second mixer is link-coupled, through L6-L7, to match the impedance of the crystal filter.

The 9-MHz Filter

The KVG Model XF-9B (See Parts list, Fig. 2) filter was chosen for a variety of reasons. It directly controls the received selectivity, and having nonlinearities in the transmitter power stages, the ultimate bandwidth and carrier suppression of the emitted signal. Obviously this is not a circuit in which to cut corners with a second-rate filter. Salient features of the 8-pole 9-MHz filter include a bandwidth of 2.4 kHz at 6 dB down, a shape factor of 1.8 at the 60-dB point, and 100 dB of stop-band attenuation, with less than 2 dB of ripple when properly terminated.

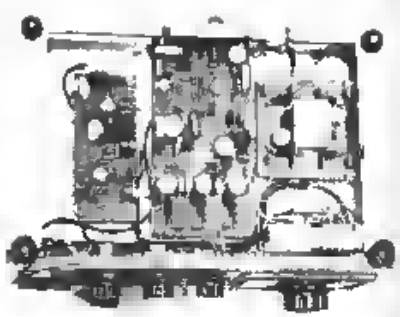
The filter does double duty in the transmitter and receiver, so diode switching was used to simplify the circuitry. Care must be exercised to avoid unwanted stray coupling between the two filter ports, as it could degrade filter characteristics. The two 45-pF trimmers, C8 and C9 in Fig. 3, were intended to provide ripple-bandpass tuning for precise adjustment of the filter response. In practice, they may be omitted, as it is doubtful that any noticeable improvement is made by their use.

Receiver I-F and Product Detector

In the XRCEIVE mode the output of the KVG filter is diode-switched to the input transformer T8-L9 in Fig. 3. This has a capacitive voltage divider, C11-C12, across its secondary, for optimum impedance match between the filter and the i-f amplifier, a CA3028A differential-amplifier IC, U1. As with other tuned stages, the 9-MHz i-f circuits are toroidal, to lessen undesired interstage coupling and attendant instability. The i-f gain is approximately 25 to 30 dB. The i-f output is sampled through an 8-pF capacitor, to provide energy for the age amplifier, Fig. 4. The control voltage is fed back to pin 7 of the i-f amplifier IC to produce about 25 dB of age dynamic range. The age voltage varies from less than 2 V at maximum gain, to 12 V during periods of age action and maximum i-f gain. Another capacitive voltage divider, C 4 and C15, provides impedance transformation between the i-f output and the product-detector input.

Of all circuits considered for the product detector, the MC1496G IC, U2, proved to be the best candidate. With 12 dB of conversion gain, the 1496 compensates for the relatively low gain of the receiver front end. It has an effective dynamic range of 90 dB, and can handle a wide variation of signal levels, despite limited age action. BFO injection at 9 MHz is generated by a 2N3904 crystal oscillator, Q9 in Part II of the six-part transceiver assembly. An injection level of 200 mV pk-pk is needed. Extreme care must be used in laying out

Bottom view of the transceiver, showing the i-f amplifier and product detector (U1 and U2) at the left side, the sideband generator (Q8) assembly at the center, and the 9-MHz filter (Z1 or Z2) at right.



and wiring the transceiver. To be sure that stray RFO energy does not reach the rf or i-f stages. Linkage at 9 MHz can kill the rf gain through the agc and all the BFO energy reaches the mixers is possible for the fourth harmonic to beat with the 36-MHz oscillator.

Audio Amplifier

A single MFC8010 or HEP C60041 is the entire speaker subsystem shown at the bottom of Fig. 3E. It is an economical plastic-case unit designed for consumer electronics use, capable of one watt continuous rms output to a 16-ohm load, with an 1dBh at 10 mV input. This much sensitivity is not needed because of the high output level of the product detector and it caused some instability as a result of stray coupling on our audio board. The final circuit details shown are for an input sensitivity of 400 mV. The amplifier is capable of driving loads of less than 16 ohms, but with a supply voltage of 15 the device simply may be exceeded and operation at high audio levels is best avoided. The stabilization network recommended in the manufacturer's application notes was ineffective in curing instability occurring with certain loads I tried. It was found that a 1-µF capacitor from pin 8 to ground would eliminate regenerative tendencies in the amplifier.

A feedback path between the product detector and the audio stages through the supply wiring, was found when certain high-impedance voltage sources (such as carbon-pile regulator) were used. This was cured with a 2000-µF electrolytic capacitor from the supply line to ground. The capacitor, shown at the control-panel diagram, Fig. 10, Part II, also helps to reduce ripple from an ac supply, at least a rectifier.

AGC

The agc system is a variation of one designed by WB1ER, for use with the CA3028.[1] Original plans were for an agc system using the MFC4040 electronic attenuator. This scheme did not work as well as desired, so the i-f agc arrangement was pressed into service. The CA3028A i-f amplifier, U4, is the only stage controlled, so the agc range is somewhat limited. Performance is not spectacular, but the agc is capable of handling the strong local signals encountered in 50-MHz operation. Part of the 9-MHz i-f output is fed to a CA3028A cascode agc amplifier, U4 in Fig. 4, producing about 40 dB of gain. Output of this amplifier is detected in a simple voltage doubler, with RC time constants selectable at this point for fast or slow agc action, via S1. The dc level from the voltage doubler is stepped up in a two-stage dc amplifier, Q3 and Q4, which supplies both S-meter current and agc voltage to the i-f amp.

25-kHz Calibrator

The calibration oscillator assembly mounted to the block diagram is a recommended built-in accessory for the transceiver. A 100-kHz crystal, Y1 in Fig. 5, is used in a 7474 oscillator. Calibration intervals of 100 kHz are of little use in the small tuning range of the transceiver, so two J-K flip-flops, U5 and U6, were added to divide the 100-kHz frequency down to 25 kHz. The rich harmonic content of the flip-flops output will mix the 50-MHz range. The calibrator output is brought into the receiver through miniature 50-ohm coax (RG174/U). Strip coupling between an exposed section of the inner conductor one inch long, and the input transformer of the receiver, gives adequate calibrator signal to the receiver. Coupling is not critical and the exposed lead can routely be adjacent to the input circuit.

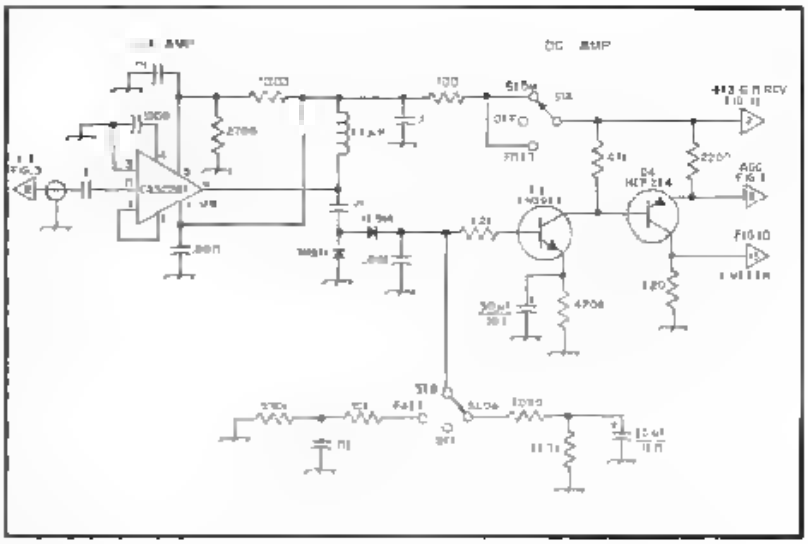


Fig. 4 — Schematic diagram of the agc and S meter amplifier. Parts not described are numbered for text reference.
S1 — 2 pole 3-position toggle switch (center off). U4 — i-f amplifier IC (RCA CA3028A).

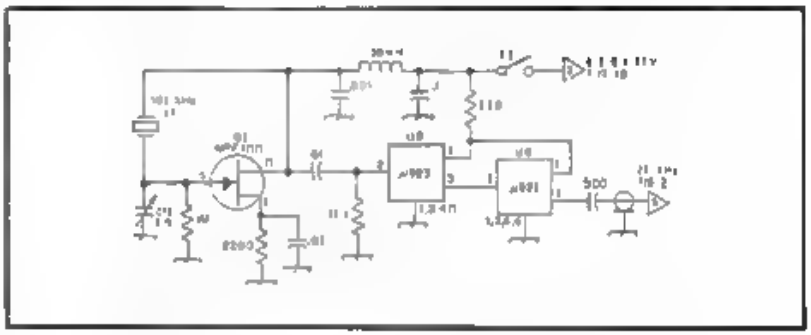


Fig. 5 — Schematic diagram of the 25-kHz crystal calibrator.

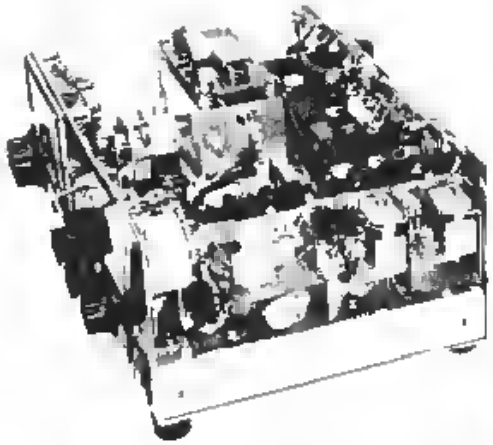
C36 — Ceramic trimmer, 7- through 45-pF.
S2 — Spst momentary on push-button.
U5, U6 — J-K flip-flop (Fairchild µ923).
Y1 — 100-kHz crystal, fo 32 pF load (International Crystal Mfg Co).

[1] See *Radio Amateur's Handbook*, 1973 and 1974 Editions, "A Receiving Package for 28 to 144 MHz," Chapter 8.

From March 1975 QST, p 27:

QRP Transceiver for 50 MHz

Part II



Side view of the KI2JH 50-MHz transceiver with cover removed, showing the transmitter rf amplifier stages, not completely seen in the top view, Part I.

IN MOPAR a practical description of the KI2JH 50-MHz transceiver was divided to put receiving assemblies in Part I and transmitting units in Part II. Several board assemblies serve both functions, including the 9-MHz filter in Part I. This rounding portion covers other dual-purpose units of the transceiver, as well as those used for transmitting only. As in Part I, order of appearance is indicated by unit numbers in the block diagram, Fig. 1.

The 5-MHz VFO

The VFO and the 36-MHz crystal oscillator, Fig. 6, determine both the transmitting and receiving frequencies. The VFO design was adapted from one used by W1KLA in a 20-meter ssb transmitter.[1] The Colpitts oscillator uses an RCA 40841 MOSFET, Q5, with gate tied together. A 1N914 stabilizes the gate voltage, reducing harmonic generation by limiting device transconductance on positive voltage peaks. Two medium-gain 2N2222A transistors, Q6 and Q7, in an emitter-follower circuit match the VFO output to 50 ohms. The 5-MHz energy passes through a low-pass filter, to eliminate harmonic output from the buffers. It then goes to the transmitting and receiving mixers through RG-174/U coax.

Oscillator voltage regulation was not found necessary, but can be done with a 9.1-volt Zener diode from the oscillator drain to ground. This diode and the 220-ohm resistor already in the drain lead should not be in physical proximity to frequency determining circuits of the VFO, as heat disruption by the resistor and diode might cause oscillator drift. A diode switched capacitance (CR19 and C20 in series) offsets the VFO during xmit operation, to preserve dial calibration.

The VFO is set to cover 5.05 to 5.19 MHz, giving band coverage of 50.05 to 50.19 MHz. That allows operation in the upper portion of the cw-only band (50.0 to 50.1), and also cw or ssb in the first 90 kHz of the ham band. Most current operation in both modes is in this range. More coverage can be had by using a larger capacitor (more plates) for C14. Note that the ceramic padder, C21, has a negative temperature coefficient. Its "N" value can be selected experimentally, if the N330 specified does not correct oscillator drift adequately. Or, a smaller capacitor of the more common N750 type, used in parallel with an NPO, can be tried, to correct drift in environments different from that in which the author's rig operates.

The VFO dial is an old National Type WCN, with a Jackson Brothers ball-bearing vernier drive in place of the original planetary mechanism. This gave the smooth free feel that is so important in a dial that is to be used for receiver tuning, as well as for VFO control. No penking or tracking controls are needed, as the limited tuning range of the equipment eliminates tracking problems.

Except for the main tuning capacitor, C15, and its padder, C17, both of which are visible in the top view, lower left corner, the VFO assembly cannot be seen in the photographs. The entire board and VFO components are in a Minibox-type enclosure, for thermal and electrical isolation. The box is under the 25-kHz oscillator board, left center of the top view. It is important that the mounting of all VFO parts be rigid and special care should be given to the dial drive and variable capacitor assembly. If necessary, the front panel and chassis can be reinforced with additional thickness of aluminum, in these areas.

36-MHz Heterodyne Oscillator

This oscillator, shown schematically in the lower portion of Fig. 6, also serves both transmitting and receiving functions. A 2N3904 transistor, Q8, is used with a 36-MHz crystal, Y2, in a reliable overtone oscillator circuit, to furnish injection for both transmitting and receiving mixers. In the receiver it heterodynes the signal to the 14-MHz range in the first mixer Q1, Fig. 2. In transmitting it beats with the 14-MHz output of the second transmitting mixer, Q13, Fig. 8, to produce the 50-MHz signal that is fed in the amplifier stages of Fig. 9.

The oscillator collector voltage is Zener-diode regulated, to maintain frequency stability and rf output level with varying supply voltage. Two lightly coupled tuned circuits are used to reduce harmonic content in the output. The trimmer capacitor C23 is adjusted to provide the minimum coupling needed to develop 1.5 volts pk-pk for the mixers. The second tuned circuit L13-C24 was added after the 36-MHz oscillator board was made. It is seen in the upper central portion of the top-view photograph, above the oscillator assembly.

Sideband Generation

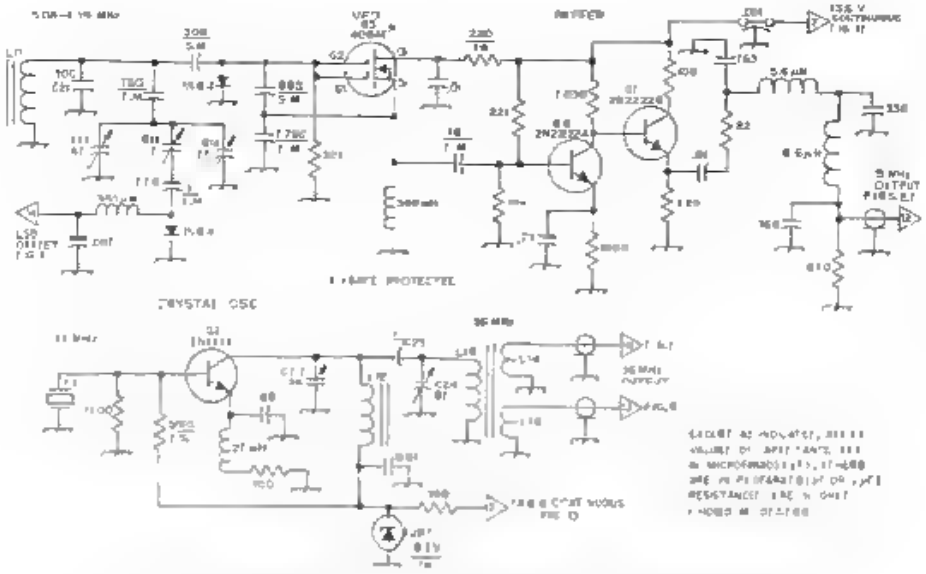
The sideband generator, Fig. 7, is the largest and most complex subassembly in the transceiver. It is seen in the middle portion of the bottom-view photograph. Included are a crystal oscillator, Q6, two speech stages, Q10 and Q11, and a balanced modulator, U7. The three crystals, Y3, Y4, and Y5, are available for use with the KVG filter, and their frequencies are related to provide upper and lower sidebands and cw, while maintaining the dial calibration for all modes.

The balanced modulator is a Motorola MC1596G IC using information supplied by K2QWR[2] BFO injection is critical for maximum carrier suppression. There should be 150 mV developed at Pin 8 of U7. This can be adjusted by substituting other values for the 82-pF coupling capacitor, C29. A multiturn Trimpot was used for the carrier balance control, R3, for accurate and stable carrier nulling. Since there is rf energy at the point, only at dc level, the control can be located away from the balanced-modulator circuitry at any convenient point, with no deleterious effects.

In the cw mode the carrier suppression is purposely unbalanced, to generate a carrier at the MC1496G output, and also a crystal within the filter passband is used. Operation with as much as possible, though not shown here, by increasing the value of the 1000-ohm resistor, R5, used for carrier insertion. A point will be found where satisfactory cw is generated, minus one sideband, of course. The two frequency-setting capacitors, C25 and C26, associated with the two sideband BFO crystals are used in conjunction with R3 in nulling the carrier. All settings interact, and it will be necessary to repeat adjustments several times, for maximum carrier rejection on both upper and

[1] *The Radio Amateur's Handbook* "A 20-Meter SSB Sideband Transmitter," 1973 Edition, p. 414.

[2] Rehnke, "Integrated Circuit Balanced Modulator," *Ham Radio* September, 1970



NOTE Enameled wire may be used wherever No 26 Teflon wire is specified in this project, so long as the builder is careful to avoid abrasion damage during construction

Fig 6 — Schematic diagrams of the VFO and 30 MHz oscillator used in the KI2JH 50-MHz noncrystal transceiver. Parts not described are numbered for text reference.
C17, C24 — Ceramic trimmer, 7 to 45 pF.
C18 — Shaft-type miniature variable, 35 pF.
C19 — Ceramic trimmer, 1.5 — 7 pF.
C20 — 1 pF silver mica.
C21 — 100 pF disk ceramic, neg. temp. coef. Select N value for maximum oscillator drift. N330 used.
C22 — Ceramic trimmer, 5 — 25 pF.
C23 — Gimmick capacitor, two 1 inch lengths No. 24 Teflon wire, twisted to give desired maximum spectrum level, 1.5 V pk-pk.
L11 — 26 turns No. 28 Teflon wire on 1/2 inch toroid core (Amidon T-50-2, red).
L12, L13 — 12 turns like L11, except Amidon T-50-6 yellow.
L14, L15 — 1 turn each, bifilar-wound at low-Z end of L13.
VR1 — 9.1-volt 5-watt Zener diode.

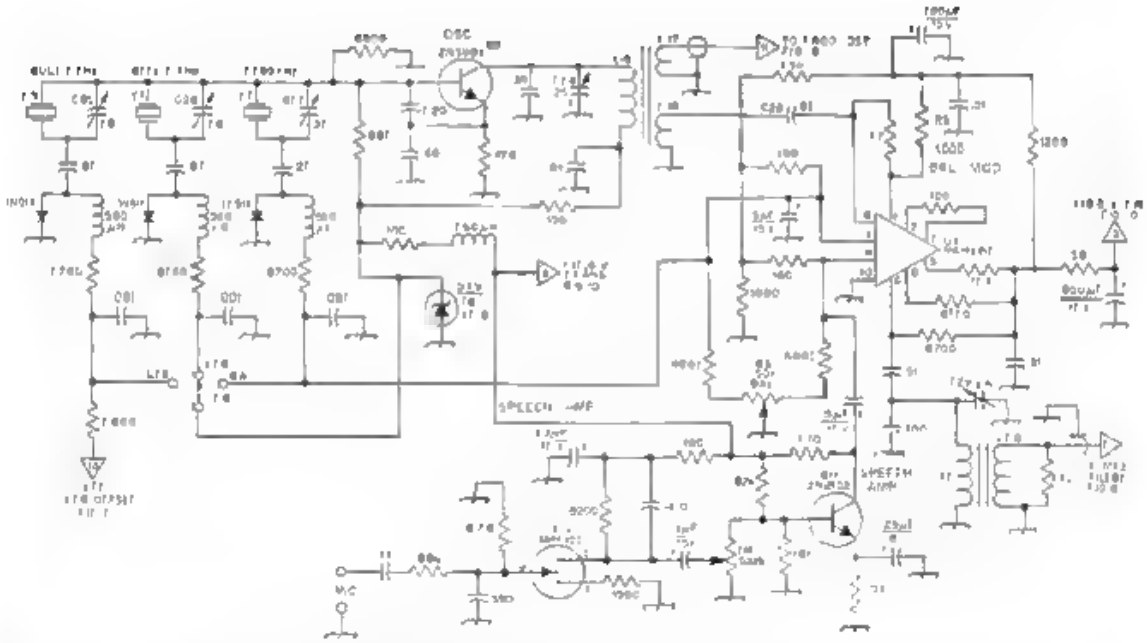


Fig 7 — Schematic diagram for the wideband generator in the 50-MHz transceiver. Parts not described are numbered for text reference.
C25-C28, incl — Ceramic trimmer, 5 — 25 pF.
C29 — Ceramic trimmer, 7 — 45 pF.
L16 — 32 turns No. 28 Teflon wire on 1/2-inch toroid core (Amidon T-50-2, red).
L17, L18 — 2 turns each, bifilar-wound at low-Z end of L16.
L19 — 20 turns like L16.
L20 — 9 turns No. 28, on core of L19.
R3 — 50,000-ohm control.
R4 — 1 Megohm audio control.
S3 — Single-pole 3-position rotary switch.
U7 — Balanced modulator IC (Motorola MC1496G).
VR2 — 9.1-volt 1-watt Zener diode.
Y3, Y4, Y5 — Crystal frequencies as indicated ordered with FL3 (Spectrum International XF902, XF901 and XF903 respectively).

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2

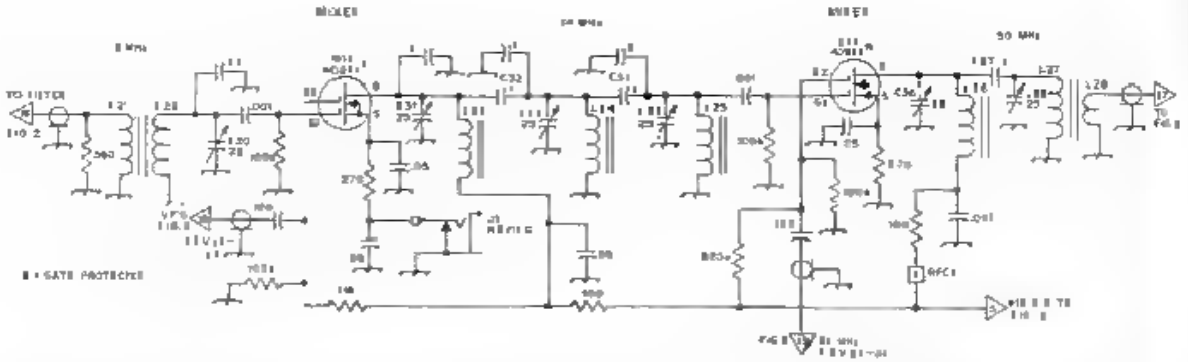


Fig. 8 — Schematic diagram of the transmitter mixer assembly for the 50-MHz transceiver.
C30, C31, C33, C35, C36, C38 — Ceramic trimmer, 5 – 25 pF.
C32, C34, C37 — Gimmick capacitor, two 1-inch lengths No. 26 Teflon wire twisted three times, approx 1 pF.
J1 — Closed-circuit jack.
L21 — 10 turns No. 26 Teflon, on 1/2-inch toroid core (Amidon T-50-2, red).
L22 — 30 turns like L21, on same core.
L23 — 27 turns like L21.
L24, L25 — 30 turns, like L21.
L26, L27 — 8 turns, like L21.
L28 — 2 turns on same core as L27.
Q12, Q13 — Gate-protected MOSFET (RCA 40841).
RFC1 — 4 ferrite beads (Amidon 43-101).

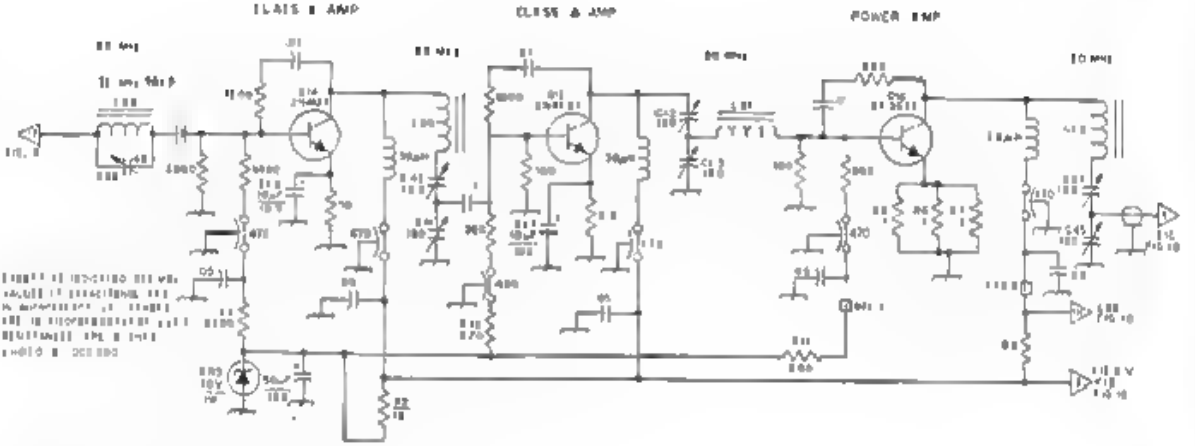


Fig. 9 — Schematic diagram of the transmitter amplifier stages.
C39 Ceramic trimmer, 7 – 45 pF.
C40 C45 (all) — Mica trimmer, 5 – 180 pF (Hamtronics 9821).
C48, C47 — Small aluminum electrolytic, 10 µF, 10 volts (Mallory MTP 106M010 P10).
L29 — 10 turns No. 26 Teflon on 1/2 inch toroid core (Amidon T-50-2, red).
L30 — 6 turns like L29.
L31 — 8 turns, like L29.
L32 — 9 turns, like L29.
Q14, Q18 — 2N4427 (RCA). 2N3866 also usable in Q15.
Q16 — 2N3633 (output). 2N3375 also may be usable.
R8, R6, R7 — 7-ohm 1/2 watt composition resistors in parallel.
R8, R10 — Approximate values. Adjust for collector currents of 60 and 100 mA for Q14 and Q15 respectively.
R11 — Approximate value. Adjust for idling current of 30 mA for Q16 or Q18 respectively (two-tone test).
RFC2, RFC3 — 4 ferrite beads (Amidon 43-101).
VR5 — 10-volt 1 watt Zener diode.

lower sideband. The trimmer C27, for the crystal IFO crystal is simply set at the point where no further increase in carrier output is obtained.

The output of the balanced modulator goes to the KVG filter where the unwanted sideband and any trace of carrier are removed. Resistive loading of the KVG ports at the *transmit* mode impedance (across L20 and L21) was done for impedance matching.

Transmitter Mixers

The transmitting converters bring the 9-MHz signal up to 50 MHz in the opposite order of conversion to that of the receive section. Both mixers, Q12 and Q13, are 40841 MOSFETs. The first up-converts the 9-MHz filter output to a 14 MHz bandpass i-f, by heterodyning it with the 5-MHz VFF output. Three lightly coupled tuned stages follow this mixer, forming a 14-MHz bandpass filter. This is very important, reducing spurious mixing products that would otherwise be compounded in the next mixer, and also eliminating harmonic multiples of the VFO frequency. (The tenth harmonic could be especially troublesome.) The source ground return of the first mixer is keyed for low operation, via J1 in Fig. 8.

The second transmitter mixer combines the 14 MHz i-f with energy from the 36-MHz heterodyne oscillator, to produce the desired 50-MHz signal. Two lightly coupled 50-MHz LC stages follow the second mixer. These reduce 36 MHz and harmonic feed-through as well as any undesired mixer products present. As with the 40673 MOSFETs, 1.5 volts pk-pk is the recommended injection level at Gate 2 of the 40841.

Amplifier Chain

Three stages are needed to bring the 50-MHz output up to the 3-watt level. The first two are 2N4427s, biased for Class-A operation. Heat sinks are needed to dissipate the heat generated in

Class-A service. The 2N4427 is designed for that service, and it is fairly "hot" at 50 MHz. Much empirical design went into the development of a stable circuit, resulting in quite a few gray hairs for the author. Collector-to-base degenerative feedback is used to reduce the gain of the 2N4427s, and this improves the stability and linearity of both stages. A 2N3632 orderly transistor leads along in the output amplifier stage. This larger-than-necessary type was used mainly because it is plentiful and low-priced on the surplus market[1] and it works well at 50 MHz. The final stage runs Class B, producing about 15 watts average ssb output.

The base biasing resistors of the 2N4427 stages, R9 and R10, may have to be altered slightly to obtain the desired collector currents indicated in the parts list. The idling current for the 2N3632 is 200 mA. All base bias levels are developed from a 10-volt line. Zener diode regulated to work with a linear operating region for the devices, with varying supply voltages. A two-tone test[1] will indicate proper biasing of the 2N3632 amplifier, and also the maximum average collector current for the stage, before flat-topping occurs. This was 450 mA, on voice peaks, in the author's transceiver. The 2N3632 draws 350 mA in cw operation. The two-tone test will also provide a good indication of spurious oscillations, which sometimes occur only at certain power levels, or with mismatched loads.

Collector current in the 2N3632 is monitored across R8, the value of which is adjusted to give full-scale deflection with 500 mA with the 5-mA meter used by the author. A variable should be made by winding fine wire on a low-value resistor experimentally. The meter circuit shown also allows monitoring of the supply voltage and relative signal strength.

A spectral analysis made at the ARRL Lab showed a very clean output from the transmitter. Only the second and third harmonics of the output frequency and a 90-kHz spurious signal were evident and in all cases there were better than 50 dB down, referred to the carrier.

Results

A brief summary of on-the-air performance with the transceiver may be of interest. During a recent VHF Party, 11 ARRL stations were worked from the author's home in North-Central Connecticut. These included all New England except Rhode Island, and Eastern Pennsylvania, Northern New Jersey, Eastern New York (NYL-Long Island, and Delaware. Not bad for three watts! Under normal conditions the contacts are made regularly with New York, New Hampshire and the Boston area stations are not to 100 miles or more with little difficulty in Dallas and out to 200 miles or so when the conditions are good. With even this

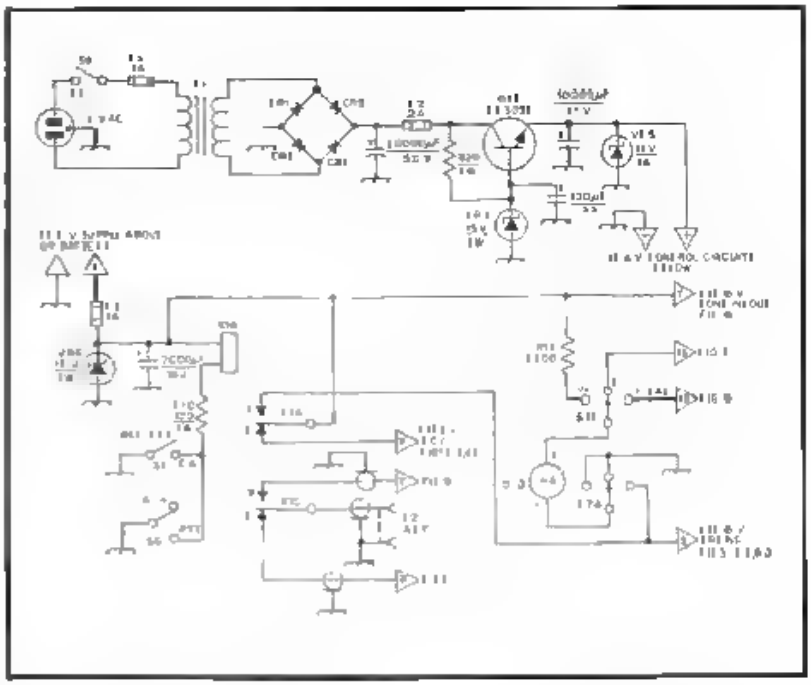


Fig. 10 – Schematic diagram of hi power supply and control circuit for the 80-MHz transceiver. Capacitors are electrolytic, values in µF.

CR1-CR4, incl – 100 PRV, 1 A
P1, P2 – 1-8 ohm
F2 – 2 A slow-blow.
J2 – Coaxial jack, SO-239.
K1A – K1C, – Miniature 2-pole, 2-position relay, 120-ohm coil. Vary wiring of, to eliminate R12 for other coil resistance.
S4, S5 – Spst toggle
S6 – PTT switch on microphone
S7 – Double pole 3 position wafer switch
R13 – 8200 ohms, or value to give full-scale deflection with meter used
VR4 – 15-volt 1 watt Zener diode
VR5, VR6 – 10-volt 1-watt Zener diode

same set-up during contests which should help to dispel concern about low power and normal communications. When the band opens for sporadic-E skip, it is only a matter of finding a clear space to jump into – and there's usually plenty of room above about 50.125 mHz. Almost anyone heard will can be worked with a little perseverance, skill, and old-fashioned luck. Locals are often skeptical of the claimed low-power level, but running this rig has been a great boon to neighborhood TV reception.

As is likely to be the case with amateur construction projects, this unit generated ideas for improvements and accessories. A recent addition is a 12-volt wide-band amplifier, capable of delivering up to 30 watts. A matching high-current power supply has also been added. A 2-meter transverter has been completed and is now in use at K1ZJH.

[1]Deleted

[1a] *Radio Amateur's Handbook*, Two-Tone Test scope patterns and gravitation, pages 401 and 427, 1973 and 1974 Editions.

From November 1986 QST, p 43:

Audio-Filter Connections For The Ten-Tec Argonaut Transceiver

☐ I have a Ten-Tec Argonaut 509 transceiver. After I purchased an outboard audio filter, there were some unexpected interface problems. When a sharply tuned filter is placed in the audio line, the sidetone can be filtered out.[1] Also, my filter supplies only about 1 W of audio, which is plenty for headphones, but not for a noisy room.

Some investigation reveals that the 509 audio is generated and preamplified on the IF board. As a matter of fact, the optional Ten-Tec filter is connected within that stage via pins 4 and 5 (FILTER OUT and FILTER IN, respectively) of the rear-apron accessory jack. (These pins are shorted when no filter is used.) There is a mix-up in the nomenclature between the IF-board schematic and the block diagram in my instruction manual. Pin 4 should be connected to the filter output, with pin 5 connected to the filter input.

My onboard filter has a fixed gain of about , which is ideal for use in the Argonaut IF stage. If your filter has some gain, the filter amplifier (usually an LM386) can be thought of as an output amplifier that is also capable of higher output power. It will probably work just as well with lower drive levels. Therefore, when placed between pins 4 and 5, most any audio filter should work.[2]

This set up leaves the sidetone and the audio output power much the same as before the filter was connected. In addition, the filter is within the Argonaut AGC loop. I have had no problems with this arrangement.—*Michael Martin, KD4ZF, 227 Neville Cir NE, Palm Bay, FL 32907*

[1]The sidetone frequency of the Argonaut 509 is adjustable and therefore does not necessarily correspond to the receive offset.—Ed.

[2]Editor's Note: If your filter has a gain control, set it for unity gain. The chief hazard here is that the filter will overdrive the 509 audio stages. If there is such a problem, simply build an attenuator to follow the filter. Part values for both T- and pi-network attenuators are given in Chapter 25 of the 1985 and 1986 ARRL *Handbooks*.

From October 1988 QST, p 40:

Curing Mechanically Induced Frequency Jumps In The Ten-Tec Argosy 525

☐ If you push with a finger on the top of the panel or case of an Argosy 525, the frequency of the rig's permeability-tuned oscillator (PTO) may change by 200 Hz or more, seldom returning to the original frequency. Here's how I eliminated this problem in my '525.

Remove the rig's bottom cover. Careful! The speaker leads are not very long and have no strain relief, so take care not to pull the leads out of the speaker. Check the left front foot screw for excessive length; mine was digging into the plastic portion of the '525's phone jack. Pressure on the ends of the front panel results in pressure on this screw, the resultant panel twist is coupled to the PTO. If this condition is present in your rig, snip off the end of the screw with cutters.

With the '525's cover removed, I discovered that touching the PTO cover or bringing part of the '525's bottom cover near the PTO cover can cause wide frequency changes. This suggests that the PTO shielding is inadequate. To correct this condition:

1) Remove the small bracket on the side of the PTO housing that normally receives one of the mounting screws for the transceiver bottom cover.

2) Loosen the PTO cover by backing the PTO-cover retaining screw out a few turns.

3) Remove the piece of fiber board that insulates the PTO cover from the PTO aluminum housing.

4) Cut a piece of household aluminum foil a little wider than the length of the fiber board and about 10 inches long. Wrap the fiber board with about three thicknesses of foil. Cut a hole in the foil corresponding to the hole in the fiber board to permit access to the PTO alignment coil slug.

5) Slip the foil-wrapped fiber board back into its original position and tighten the PTO-cover retaining screw to clamp the foil to the PTO housing.

6) Reinstall the bottom cover of the '525, omitting the screw that formerly engaged the PTO bracket.

This completes the modification. Note: This procedure shifts the Argosy 525's tuning calibration somewhat, so you may need to reset the tuning dial to restore proper calibration. If you find that the tuning shift is excessive or the dial tracking is off, consider realigning the PTO as described in the '525's manual.—*Charles J. Michaels, W7XC, 13431 N 24th Ave, Phoenix, AZ 85029*

From February 1986 QST, p 51

AGC and RF-Gain Controls for the Ten-Tec Argosy

☐ I have met many users of the Ten-Tec Argosy transceiver on the air and, while all agree the rig is a fine performer, most wish it had an RF-gain control. Since the Argosy operates QSK and uses only AGC to set the RF gain, the noise between dots and dashes can be quite raucous, as the receiver gain is wide open until the AGC takes control. My outboard RF gain control requires absolutely no surgery to the rig and is within the ability of nearly anyone; the only disassembly required is removal of the top cover.

The circuit in Fig 1 applies an adjustable voltage to pin 5 of U1 (MC1350), which is the AGC input. It controls RF gain in the same fashion as the AGC and has no effect on normal AGC operation. S-meter readings decrease along with the RF gain. Place the outboard control on any breadboard, box or what have you. Connect the control to the transceiver by passing wires through the centers of the rivets that secure the phono-jack panel to the transceiver rear panel. The ground wire of the new control is connected to the ground wire of the jacks, just inside the '525 rear panel. Obtain +12-V dc in the same manner from the 12-V AUX jack inside the rear panel.

To make the control lead, slip a ferrite bead over the diode lead (cathode), and form the shortest hook with which you can work. Solder the hook to the lead of R29 (10 kΩ), which is centered on the end of U1.

My control works nearly as well without the diode and ferrite bead, but I seemed to get a bit of filter blowby without them. This RF-gain control definitely improves CW operating convenience, especially on a noisy band.—*Ned B. Smith, N8CWW, RR 1, Box 163, Ryan, IA 52330*

☐ Ten-Tec issued a bulletin, TN2-525, describing how to install an RF-gain control in the Argosy 525. It implies that a small, concentric, dual-10-kΩ potentiometer be installed in the AF-gain position. Such a "pot" I have not, so I added an outboard RF-gain control and found it to be a big help.

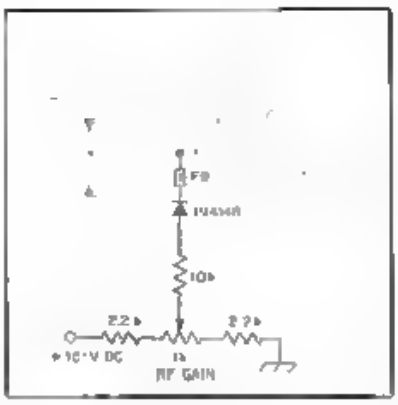


Fig 1—Schematic diagram of N8CWW's RF-gain control circuit for the Ten-Tec Argosy 525.

Then, inspiration struck: Why not reverse the Ten-Tec design and have a fixed audio gain with variable RF gain? My scheme worked well, and you can have the final version without drilling any holes. Furthermore, you can return to the original layout very easily by plugging the connectors from the original audio potentiometer back into terminal 43.

To perform the modification, proceed as follows: Remove the two connectors from terminal 43 of the IF/AF board and move them aside for future use. Wire a miniature 10-kΩ potentiometer to a four-wire connector that will plug into terminal 43. Adjust the potentiometer for an optimum audio level.

Next, turn your attention to the two connectors that are wired to the original audio-gain potentiometer and hook them up as follows: The adjustable arm of the potentiometer goes through a 1N4148 diode to the common junction of D9 and D10. Connect one mid terminal of the potentiometer to ground and the other to +12-V dc.

Set the RF-gain potentiometer to midscale and proceed with the "smoke test." I found the adjustment critical because the full range is only a couple of dial markings. (If the RF-gain control works backwards, reverse its battery and ground connections.) In spite of the RF-gain control, a strong signal still generates unwelcome audio pops, so I added an AGC ON/OFF switch. (An AGC-timing modification for the Argosy semi appeared in the November 1983 Hints and Kinks column. —Ed.)

The AGC ON/OFF switch was created by breaking the connection between D9 and Q5, and wiring in a switch. Disconnect the ac leads from the switch associated with the new RF-gain control and use that switch as the AGC ON/OFF control. (This leaves the radio without a power switch. When used with a switched supply, such as the Ten-Tec 225, the ac-switch leads may be connected, in which case power to the '525 is controlled by the switch in the power supply. When a battery or unswitched supply is used, place a power switch (35 V, 9 A) in the dc line to the radio, or add a new switch to the '525 in a location of your own choosing.—Ed.)

Full QSK CW operation is a most satisfactory experience using a manual RF-gain control and no AGC. SSB works well with the AGC on.—*Jack L. West, W6VJD, 3670 Montclire St, Sacramento, CA 95821*

From January 1984 QST, p 30:

Some Practical Antenna Considerations

City lot or "rancho grande," DX or stateside communication, we need certain types of antennas to match available space and operating preferences.

By Doug DeMaw,* W1FB

I remember the mess I made of things back when I erected my first ham antenna. Nobody told me it wasn't just a matter of erecting a wire of a specific length (130 feet was the magic number I'd picked up for 80 through 10 meters back then). Somehow, I had failed to learn that the end-fed wire had to be matched to the transmitter, and that the height above ground had a lot to do with how far away my signal could be heard. Perhaps some fundamental knowledge can save you the agonies that many of us had to endure at the start of our ham radio careers.

As I look back on that first installation at WN8HHS (Novice), I recollect the nail biting, finger drumming and the staring into space that came as a result of being unable to make my homemade CW transmitter develop output power with that end-fed wire attached to it. My first week on the air netted a handful of contacts on 80 meters — none of which were over paths greater than a few city blocks!

Then, quite by accident, the transmitter showed high PA (power amplifier) plate current at the dip (resonance), and I began to work stations all over the USA. What had changed? Earlier that day, I had added an improved manual TR (transmit-receive) switching arrangement to go from transmit back to receive (actually, it was a knife switch and some added wire in the shack). Could this have helped me? I changed things back to their original state, and sure enough — the transmitter wouldn't load up!

I learned later on that the extra feet of wire (plus the switch) I placed in the antenna line had changed the feed-point impedance of the wire, making it just right for a suitable match between the antenna and the transmitter output amplifier. Had I known about antenna tuners then, the problem would never have existed: I could have matched the wire to the transmitter and receiver for use in any of the high-frequency bands. The purpose of this article is to round off some of the sharp edges on antenna problems that could confound the beginner. The topics are based on oft-repeated questions we've answered at ARRL Hq. over the years.†

What Kind of Wire Is Best?

You'd be surprised to know that a great number of hams — new and experienced — are uncertain about which type of wire is best for antenna work. "Will insulated wire be okay?" Another query has been, "Will aluminum or steel wire radiate satisfactorily?" as well as "What wire diameter (gauge) must I use?" Well, the straight dope is that none of these are especially critical when you are dealing with wire types of antennas below VHF. If I were to offer a rule of thumb for these questions, I'd say something like, "Use whatever you can load up quickly and inexpensively." Of course, the strength of the wire should be sufficient to provide longevity and safety.

The Matter of Insulation

I'll always remember the amateurs who asked me if they could use antenna wire covered with plastic insulation. Perhaps it is a reasonable thing to wonder about; after all, insulation is an electrical barrier at dc (direct current) and can be a barrier in some ac (alternating current) circuits. Despite this, I have used all manner of insulated wire in my antenna systems, and most of them have worked quite well. Among the wire types employed were nos. 12 and 14 solid and stranded house wiring with plastic jacketing, ordinary electrical hookup wire, cotton-covered bell wire, pieces of ac line cord and, of course, enameled or Formvar® -insulated copper wire.

The insulation does not impair the radiation properties of the antenna. In fact, I prefer insulated wire, because it virtually prevents unwanted oxidation of the copper or aluminum conductor. In some cases it adds strength to the wire — another benefit.

The classic antenna wire among beginners seems to be the stranded bare copper that can be obtained at many parts stores. This is acceptable wire, but it will turn black or green rather quickly in polluted air, such as we find in industrial areas. It can become brittle and break in only two or three years if the air contains con-

*ARRL Contributing Editor, P.O. Box 250, Luther, MI 49656

†Notes appear at end of article.

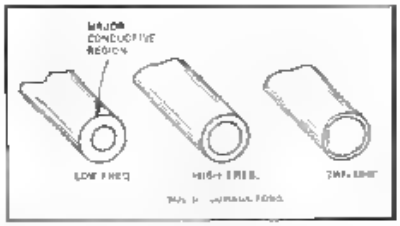


Fig 1 — Illustration of the skin effect at a conductor for various frequency ranges. The electron flow is more effective (greater penetration) as the frequency becomes lower.

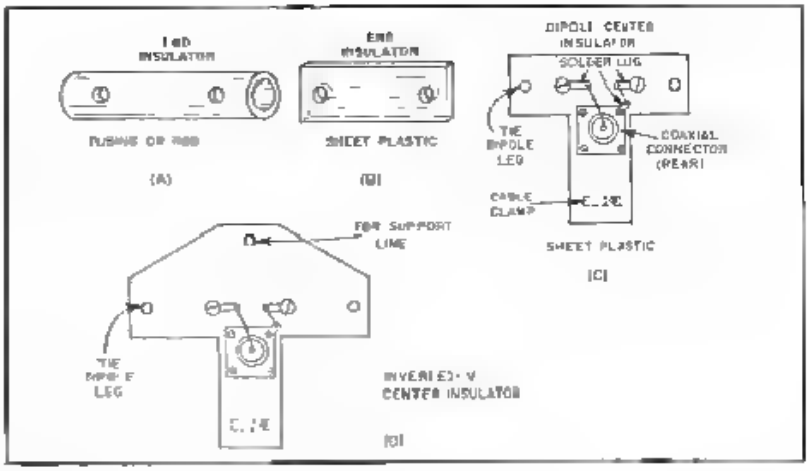


Fig 2 — Various end insulators and center blocks made from plastic material. Fabricate your own antenna hardware to save money.

siderable salt and/or acids. Frequent replacement can be costly!

If insulated wire other than the enameled type is used to prevent corrosion, be sure to seal the open ends with epoxy cement to prevent migration of pollutants and moisture into the space between the wire and the jacketing material. A marvelous new antenna wire with plastic insulation and rugged conductors was recently made available to amateurs.[1] If you are thinking of a new antenna for many years of use, this product may be of interest to you.

There may be an exception to the statement that insulation does not affect antenna performance. I was told by two experienced amateurs that they had difficulty when fashioning cubical-quad elements from vinyl-insulated house wire. The length formulas for the loop elements were of no use when using that style of wire. I haven't investigated the phenomenon yet, but the cause of the difficulty may be related to a change in the propagation factor of the wire, caused by the insulation, with the one-wavelength dimension. At VHF and higher, there is a definite difference between the propagation factor (wave velocity) of bare wire and a conductor with thick insulation when dealing with conductors that are long in terms of wavelength.[2] I have never observed velocity problems when using insulated wire in ordinary antennas for frequencies lower than 30 MHz.

Conductor Material

Can we use steel wire in our antennas? What about aluminum? Isn't copper best? Here we have to ask ourselves what is meant by the word best? That word can apply to such matters as *strength*, *weight*, *conductivity* and *cost*. If I were to ignore cost and handling convenience, and had to give but one answer, I would specify Copperweld® wire. This is a steel-center wire with an outer layer of copper. The combination provides good conductivity and strength. Most amateurs choose no. 16 gauge as a suitable "happy medium" size. But, no. 18 wire is also quite strong, and it is a little easier to work with. (Anyone who has struggled with a coil of spring-like Copperweld will understand what I mean by "easier to work with"! A loose coil can be as cooperative as a snake waiting to strike!)

Although iron and steel are not as effective a conductor at radio frequencies as are aluminum or copper, it isn't so poor that we should ignore it. I have erected a number of fine antennas with steel guy wire as the radiator elements. I have also used thin inexpensive electric-fence wire that can be purchased from Sears. A quarter-mile roll costs less than $15! Similar wire, at slightly higher cost, is available in aluminum.

The reason we may prefer good conductors to less effective ones is to reduce losses in the system. The greater the resistivity of the conductor, the greater the power loss in heating (I^2R losses). Conductivity is also based in part on the operating frequency. We have a condition that is known as "skin effect" — the ability of the RF energy to penetrate the conductor. The effective conducting area of a solid conductor is governed by frequency and skin effect (see Fig. 1). Therefore, the larger the conductor, generally speaking, the better the conductivity as the operating frequency is raised. Also, the smaller the wire diameter for a given frequency, the more restricted the antenna bandwidth, owing to increased Q (quality factor) of the system. In other words, the higher the Q of any resonant circuit, the narrower its bandwidth will be. This applies to tuned circuits, filters and the like.

I have been asked such questions as, "What is the smallest wire diameter I can use with my kilowatt rig?" If we don't consider the fragility of very small wire, we might say that even no. 28 wire can be used. I've used no. 24 and no. 26 enameled wire a number of times in so-called "invisible antennas" that were configured as end-fed random-length wires. I have yet to burn up a small-diameter wire used in that manner. The CW or SSB duty cycle, plus the air cooling of the wire, prevents current from burning up the conductor. Small-diameter wire also works nicely in radial systems (buried or above-ground systems of wires that serve as a ground screen for antennas).

Aluminum wire, such as clothesline or electric fencing, is also satisfactory for antennas. The two problems we may encounter are (1) difficulty making a good electrical joint and (2) crystallization of the wire with stress and time, which causes breakage. The use of aluminum wire generally requires the mating of copper to aluminum somewhere along the way, and this invites the rapid oxidation that is so common when dissimilar metals are joined.

Some hams have been fooled by fate when they erected antennas made from soft-drawn copper. Magnet wire, such as we wind coils from, is a form of soft-drawn copper. Although it is easy to work with since it is not prone to kinking easily, it does stretch under stress.

The longer the antenna, the more pronounced the effect. If the low SWR point in your system has changed mysteriously chances are your dipole or other wire antenna has become longer as a result of wire stretch. If this happens, you will have to readjust the system by trimming off the excess wire. Soft-drawn copper wire with vinyl jacketing is less likely to change dimension from weight, wind and icing stress.

Insulators

If you've priced commercial antenna insulators recently, you may have concluded (as I have) that the dies from which they are cast must be made of gold or platinum! I object to paying $2 or $3 for an item that is mass-produced from 25 cents worth of material. So, I make my own insulators when possible. Generally, we should strive to use insulators that are of high dielectric quality, such as ceramic, steatite, Teflon, polyethylene and Plexiglas. Other good materials are fiberglass, glass-epoxy circuit-

board material (copper removed), phenolic and other low-loss modern plastics. Many of these materials can be purchased as scrap at industrial-plastic outlets, or at a flea market. Fig. 2 shows some of the insulators we can fashion from insulating stock.

In the early days of Amateur Radio, it was not uncommon to find operators who were using antenna insulators made from pieces of hardwood or dowel rod. The wooden sections were cut to size, drilled, then boiled in tanning wax or beeswax until they were thoroughly treated against moisture. Spreaders for open-wire feed line were also made from impregnated wood.

Nylon cord is suitable for use as end insulators for wire antennas. Two or more feet of line should be used to ensure that losses are minimized when the line is wet from rain or dew. At this time, I am using a trap-style inverted-V that has 10 feet of strong nylon cord at each end.' The cord serves as a support and insulates the ends of the wire from the ground stakes.

Other items that enterprising hams have used as insulators are plastic clothespins, the bodies of plastic pens, plastic pill bottles, nylon center hubs from photocopy-machine paper rolls, plastic hair curlers, nylon six-pack headers and the solid polystyrene center insulation from RG-8/U coaxial cable. I once saw an antenna that had 8-inch strips of inner tube (discarded after a tire blowout) as end insulators! Since most rubber today contains a lot of impurities (such as lamp-black soot), I doubt that I'd use the material in my antenna system. But, this does point out that a little ingenuity can save us time and money.

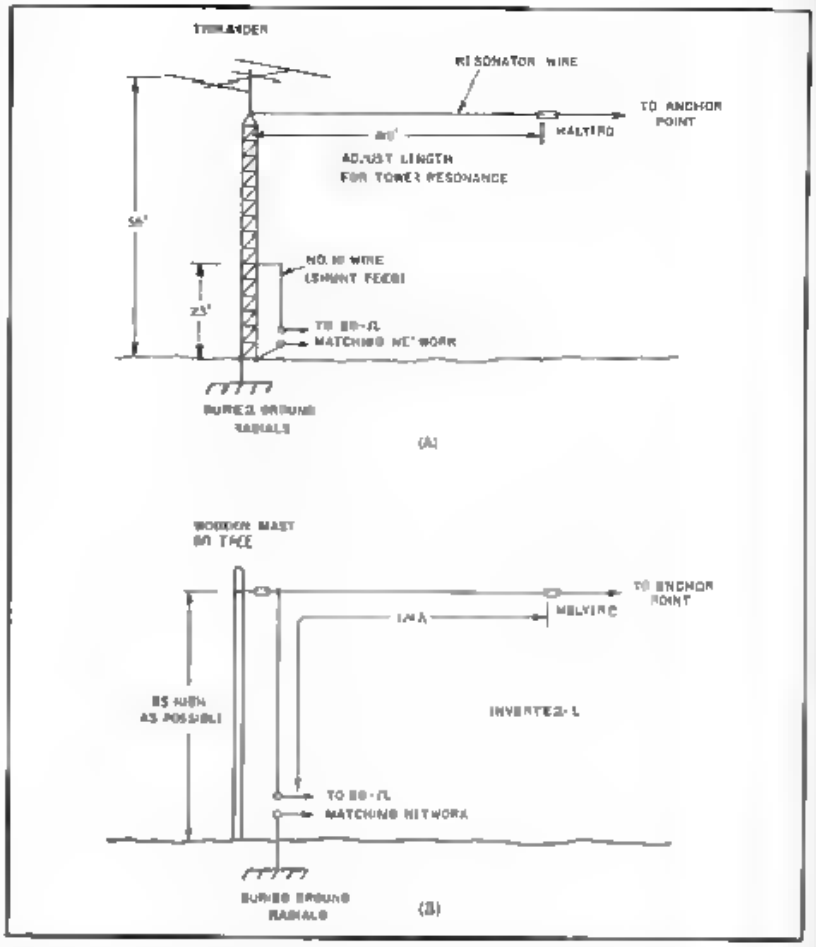


Fig. 3 — An example of how a tower and tribband Yagi might be used as a top-loaded vertical. The beam antenna provides some of the top loading, and the extender wire completes the job. The shunt arm can be spaced 2 to 3 feet from the tower. An L network or other matching system can be housed in a box and located at the feed point. This method is applicable to any unguyed tower under 120 feet in height. The shorter the tower, the longer the extender wire. If the tower is guyed, insulators should be installed in the tower connection points. Fig. 3B shows the details of a similar antenna — the inverted L.

DX or Local QSOs — Which Antenna?

The first section of this article can be considered a lengthy Hint and Kink. I hope the column editor, Larry, WA3VIL, will forgive me for my transgressions! But now that we have talked about some hardware fundamentals, what about the antenna as a whole?

All amateurs are interested in antennas, even though they may never build a piece of ham gear. There is a mystique about antennas that lures all of us. Fortunately, that is one part of radio that most amateurs will try their hands at, and the experiments can usually be carried out in a short period at a minimum outlay of cash.

But, what do we desire in terms of signal coverage? A good antenna must be designed for the distance we want to cover reliably from day to day. Some DX antennas are of little value for close-in work, and many antennas for local work are poor DX performers. Increased antenna height will enhance our DX capability, whereas the lower antennas are much better for working out to a few hundred miles in the lower portion of the HF (high frequency) spectrum. Then there's the matter of limited space for the city dweller. Many urban hams can't erect a tower, and conclude, therefore, that DX is out of reach. In this discussion, our principal concern is for high- or low-angle radiation from the antenna.

Some Easy Antennas

There is a saying among DX chasers who haunt the 160- and 80-meter bands: "A short vertical antenna and ground system is much better than a full-size horizontal antenna that is less than a half wavelength above ground." I tend to agree with that philosophy, having had the good fortune of confirming 72 countries over a three-year span on 160-meter CW. The antenna was a 50-foot, shunt-fed tower with a mediocre ground-radial system. A tribband Yagi sat atop the tower. With the same setup (and 600 W of dc input power to the last stage of my transmitter), I obtained my Worked All States Award on 160 meters. Earlier, I tried inverted Vs and low horizontal end-fed half-wave wires, but they failed miserably in DX work. They were super, however, for contacts out to a few hundred miles. The same vertical antenna was used on 80 meters with outstanding results. I had only 16 buried radials to the city-lot lawn, the longest of which was only 100 feet in length. Some were only 40 feet long. Fig. 3 shows the details of the antenna. For those who don't have a tower, a metal mast can be used in place of the tower. If only a tree is available for a support, you might try the inverted L antenna of Fig. 3B. It should provide similar results to those of the antenna at Fig. 3A.

A ground-mounted 40-meter vertical is easy to erect and is fairly "low key" with regard to being seen by neighbors. We need not use tubing if a tree support is available. A vertical wire can serve as the driven element of the antenna. Even a wire that is sloped less than 45 degrees will have

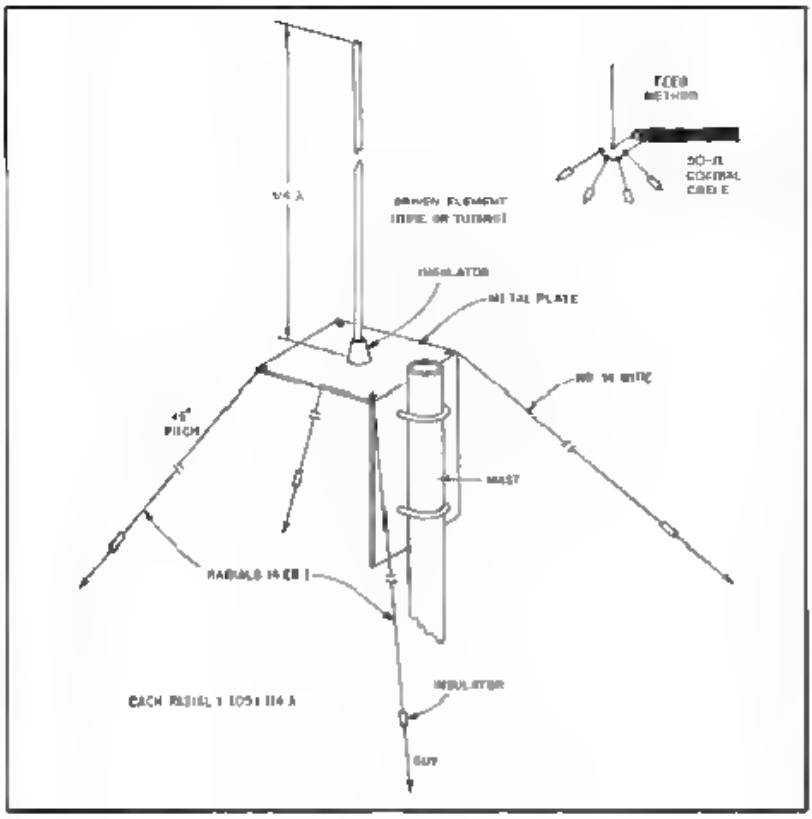


Fig 4 — Example of a ground-plane vertical. The radial wires are connected to the metal base plate and drooped at a 45-degree angle to provide an impedance match to 50-ohm line. The vertical element can be made of tubing or a wooden support can be added above the base plate to accommodate a wire element in place of the tubing. If this is done, the wire must be insulated from the wooden mast by means of standoff posts. The radial wires serve as guys for the metal system. Each wire is 5 percent longer than the driven element. This is a good DX antenna for 20, 15 or 10 meters, owing to its low radiation angle.

predominantly vertical, low angle radiation.

For operation at 20, 15 or 10 meters, it is more practical to erect a ground-plane vertical on a pipe mast or chimney mount. Four above-ground radials are sufficient for good operation. They can be made of wire and used as guy wires (see Fig. 4).

The practical limitation of low-angle vertical antennas is the inherent "dead zone" in signal coverage. Signal levels will be high within the ground-wave contour (usually under 100 miles), then there will be a skip zone where the signal is very weak (a couple of hundred miles or more) until refraction bends it down to earth beyond the dead zone. That is why many hams with vertical antennas have communications difficulties on 160, 80 and 40 meters when trying to work someone relatively close to them. A simple horizontal antenna, close to the ground, is frequently used for close-in QSOs.

A very good high-angle antenna for use on 75 or 40 meters is shown in Fig. 5. I dubbed this antenna the "Lazy Quad" when I wrote it up for *CQ* Magazine in the early 1950s. It is excellent out to, say, 500 miles — especially at those times of the day when the band is changing (near sunset and just after daylight). The ground below the antenna acts as a reflector, and the signal is directed skyward. Generally speaking, a dipole that is low to the ground has the same characteristics, and that is why it is so effective for short-haul contacts. A dipole anticone has little or no directivity

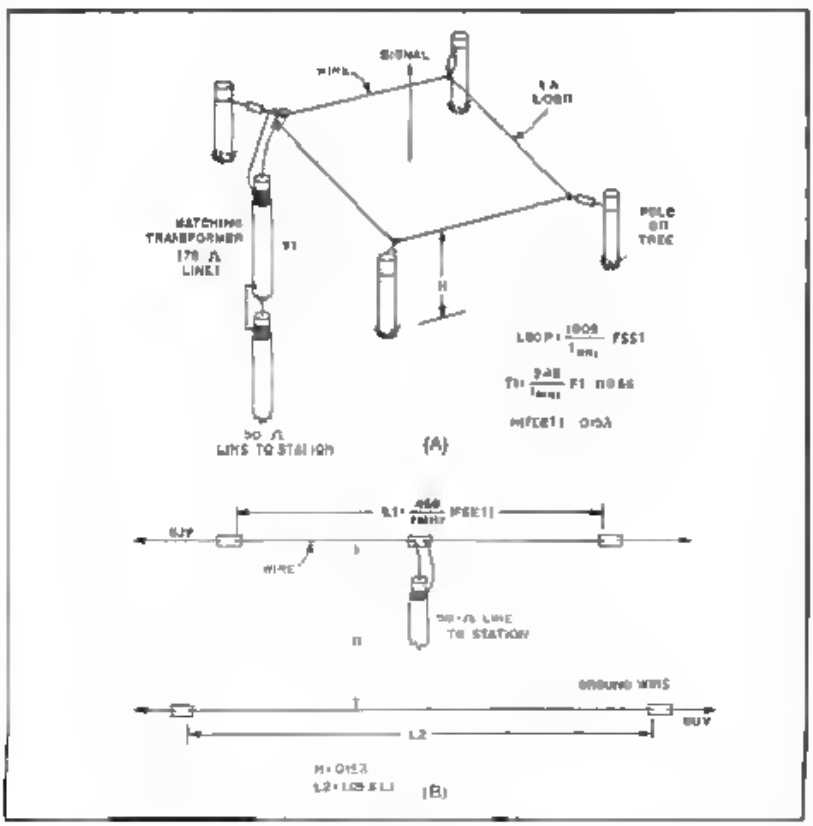


Fig 5 — The antenna at A is designed for high-angle (short range) communications on 75, 80 or 40 meters. The ground below it acts as a reflector; the better the ground conductivity, the better the performance. A coaxial transformer matches the 50-ohm feed line to the antenna. The free-space feed impedance is on the order of 115 ohms. It will be somewhat lower when so close to ground. The actual impedance will depend on the quality of the ground below and near the loop. A counterpoise loop made 5 percent longer than the driven element can be placed 0.15 wavelength below the grand loop if there is doubt about the ground conductivity in the area. A similar system is shown at B. It uses a simple dipole above a counterpoise ground or reflector. It can be used without the counterpoise ground if the earth conductivity is acceptable for skyward directivity.

unless it is a half wavelength or greater above ground. Now, that is pretty high at 160 meters (259 feet) or 80 meters (133 feet at 3.7 MHz). We hams tend to think of antenna height in terms of physical dimensions rather than electrical ones. That's a mistake, for even though 70 or 80 feet seems high, it's very low in terms of wavelength at the lower frequencies. To have an 80-meter dipole 50 feet above ground is about as poor as mounting a 10-meter beam 3 feet above ground. None of us would want to do that! It is for this reason that a short vertical antenna usually outperforms a low horizontal antenna for DXing.

We must recognize in this discussion that an electrically short antenna, vertical or horizontal, is not as efficient as a full-size antenna. There is always a trade-off to accept. Also, vertically polarized antennas are noisier during receive than are horizontal antennas. This is because most man-made noise is vertically polarized.

It would be impractical to attempt to describe the many wire antennas suitable for DX and local operation from a city lot. *The ARRL Antenna Book*, recently revised considerably, contains a wealth of practical information for those who want to build antennas. If you don't have a copy, you should invest in one.

Ground Systems in Brief

Countless amateurs have said, "I can't put up a ground-mounted vertical because I don't have room for buried radials." "Balderdash," I am prone to reply. An imperfect ground system is far better than none at all! It is surprising to observe the loud signals that some stations propagate with inferior ground screens. I remember vividly the whopping signal from W7DO[illegible]/6 when I worked 160 meters from Connecticut. He was usually the loudest station on the West Coast, and he told me he was using an 80-foot vertical with no ground radials! I dread to think about the kind of signal he would have sent my way if he had had 120 quarter-wavelength radials deployed!

Those fainlists who won't even experiment may be affected by a case of lethargy. I think experimenting is the better part of Amateur Radio. Try a vertical antenna, even if you can lay down only one or two radials. You could be rewarded with better results than theory dictates. I have always made an effort to tie as many ground wires as possible to my antenna systems. If there is a chain-link fence on your property, tie it into the ground system. Do likewise with the cold-water lines in your home, rods driven into the soil near the base of your vertical and utility-company grounds on your property.

Radial wires need not be buried in the ground. They can be laid on the lawn and staked down with homemade large staples to permit mowing the grass without hardship. If they can't be laid out linearly from the base of the antenna, wrap them around the house, garage and trees. The main idea is to get them in or on the ground — some place.

For those of you who are afraid of disfiguring your lawn by putting radial wires in it, take heart. A lawn-edging tool makes a narrow slit, and the wires need be only a couple of inches below the surface to be out of the way. The slits can be closed by stepping on them. The grass will soon grow over the incisions and no one will ever know that an "operation" took place.

What Have We Learned?

In essence, the intent of this article was to kindle your courage toward building and experimenting with antennas. Numerous cost-saving shortcuts have been presented with the hope that you will have some new tricks in your bag when you tackle that next antenna job. If you're wealthy and want to be top dog in the DX pileups, buy your antenna system. The antennas described here will make no one a "big frog in a little pond," but they'll enable you to enjoy good communications most of the time.

Notes

[1]ARRL members may take advantage of the free TIS (Technical Information Service) at Hq. by writing to the Technical Department. Limit the number of questions with each request and be sure to include a business-size s.a.s.e. for the reply to your inquiry.

[2]Deleted.

[3]J. Hall, ed., *The ARRL Antenna Book* (Newington: ARRL, 1992).

[4]1 in. = 1 x 0.3048 mm = in. x 25.4

From February 1983 QST, p 15

Antennas for Those Who Can't Have Antennas!

Radio amateurs don't engage in covert activities, but there are times in all of our lives when hidden or "invisible" antennas are necessary if we are to get on the air.

By Doug DeMaw, W1FB
ARRL Contributing Editor
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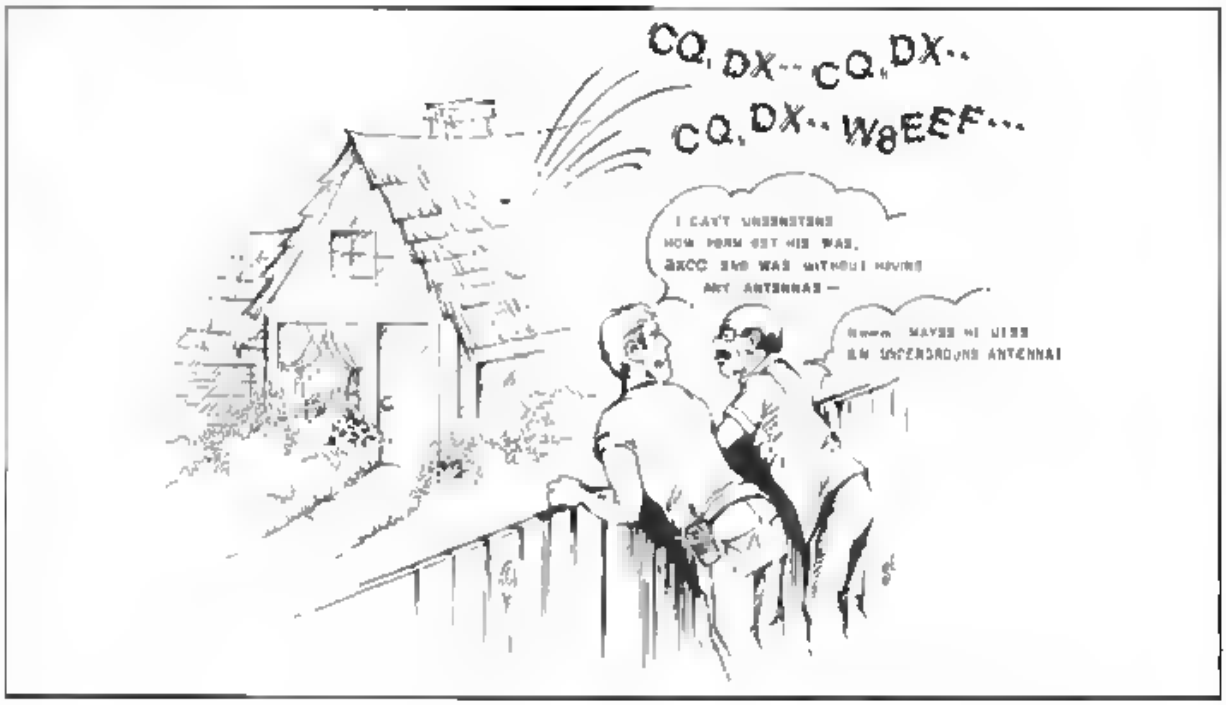

The unfortunate fact of the matter is that some radio amateurs dwell where antennas are prohibited. In other situations the operator may not want to erect outdoor antennas for fear of neighborhood opinions that he or she is destroying the beauty of the residential area. We amateurs don't regard our antennas as eyesores; in fact, we almost always regard them as works of art! But there are occasions when having an outdoor or visible antenna can present problems, especially for those who live in apartments.

When we are confronted with restrictions, self-imposed or otherwise, we can take advantage of a number of options toward getting on the air and radiating at least a moderately effective signal. In this context, a poor antenna is certainly better than no antenna at all! A number of techniques enable us to use indoor antennas or "invisible" antennas out of doors. Many of these systems will yield good to excellent results for local and DX contacts, depending on band conditions at any given time. *Don't erect any antenna that can present a hazard (physical or electrical) to humans, animals or buildings. Safety first!*

Invisible Antennas

In some areas, clotheslines are attached to pulleys (Fig. 1) so that the user can load the line and retrieve the laundry from a back porch. Laundry lines of this variety are accepted parts of the neighborhood "scenery," and can be used handily as amateur antennas by simply insulating the pulleys from their support points. This calls for the use of a conducting type of clothesline, such as heavy-gauge stranded electrical wire with Teflon or vinyl insulation. A high-quality, flexible steel cable (stranded) is suitable as a substitute if one doesn't mind cleaning it each time clothing is hung on it.

A jumper wire can be brought from one end of the line to the ham shack when the station is being operated. If a good electrical connection exists between the wire clothesline and the pulley, a permanent

connection can be made by connecting the lead-in wire between the pulley and its insulator. A Transmatch can be used to match the "invisible" random length wire to the transmitter and receiver.

Invisible "Long Wire"

In reality, an antenna is not a classic "long wire" unless it is one wavelength (or greater) long. Yet, many amateurs refer to all relatively long spans of conductor as "long wires." For the purpose of this article we will assume we have a fairly long span of wire, and refer to it as an "end-fed wire."

If we use small-diameter enameled wire for our end-fed antenna, chances are that it will be very difficult to see against the sky and neighborhood scenery. The higher the wire gauge, the more "invisible" the antenna will be. The limiting factor with very fine wire is fragility. A good compromise can be realized by using no. 24 or no. 26 magnet wire for spans up to 130 feet (m = ft × 0.3048). Lighter-gauge wire can be used for shorter spans, such as 30 or 60 feet. The major threat to the longevity of fine wire is icing; also, birds may fly into the wire and break it. Therefore, this style of antenna may require frequent service or replacement.

Fig. 2 illustrates how we might install an invisible end-fed wire. It is important that the insulators also be lacking in prominence. Tiny Plexiglas blocks work well, as do small-diameter, clear plastic medical vials. Some amateurs simply use rubber bands for end insulators, but they will deteriorate rapidly from sun and air pollutants. They are entirely adequate for short-term operation with an invisible antenna, however.

The invisibility of the antenna can be carried even farther if one is willing to use camouflaging techniques. This can be achieved by spraying the antenna wire with green, tan, brown, black and light blue paint at 1-foot intervals. In some instances, a single layer of gray or medium-blue paint will help to disguise the antenna. The wire must be free of grease and dirt if paint is applied, and the paint should be of "exterior" grade. This camouflaging effect can also be realized by dipping sections of the wire into cans of paint of the appropriate colors, assuming that spray paint is not available or desired.

Rain-Gutter or TV Antennas

A great number of amateurs have taken advantage of standard house fixtures when contriving inconspicuous antennas. A very old technique is the use of the gutter and downspout system on the building. This can be seen in Fig. 3. A lead wire is routed to the shack from one end of the gutter trough. We must assume that the wood on which the gutter is affixed is dry and of good quality in order to provide a reasonable insulation factor. The rain-gutter antenna may perform quite poorly during wet weather or when there is ice and snow on it and the house roof.

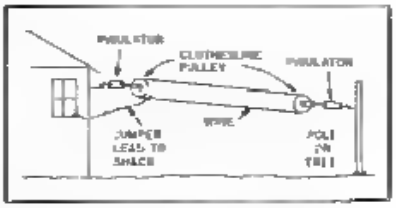

Fig. 1 — One form of hidden antenna is shown in this drawing. The antenna performs double duty as a radiator and a washline. Large-diameter insulated wire (flexible) can be used as the line, or a clothesline with a wire center may be employed. The pulleys are insulated from the house and the pole.

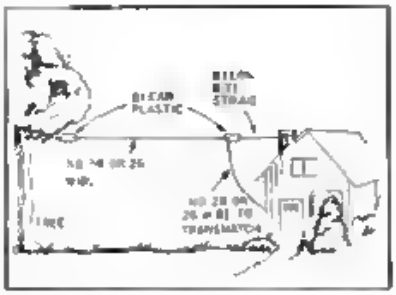

Fig. 2 — Light-gauge enamel-covered wire works really as a hard-to-see antenna. Although a bird may knock it down from time to time, reerecting it beats having no antenna at all.

We need to mention that all joints between gutter and downspout sections are bonded with straps of braid or flashing copper to provide good continuity in the system. Poor joints can cause rectification and subsequent TVI and other harmonic interference. Also, it is prudent to insert a section of plastic downspout about 8 ft above ground. This will prevent humans from receiving rf shocks or burns while the antenna is being used. Improved performance may result if the front and back gutters of the house are joined by a jumper wire to increase the area of the antenna.

Fig. 3 also shows a TV or fm antenna that can be employed as an invisible amateur antenna. Many of these antennas can be modified easily to accommodate the 144- or 220-MHz bands thereby permitting the use of the 300-Ω line as a feeder system. Some fm antennas can be used on 6 meters by adding no. 10 buss-wire extensions to the ends of the elements and adjusting them for a VSWR of 1:1. If 300-Ω line is used it will require a balun or Transmatch to interface the line with the station equipment.

For operation in the hi bands we can tie the TV- or fm-antenna feeders together at the transmitter end of the span and treat the overall system as a random-length wire. If this is done, the 300-Ω line will

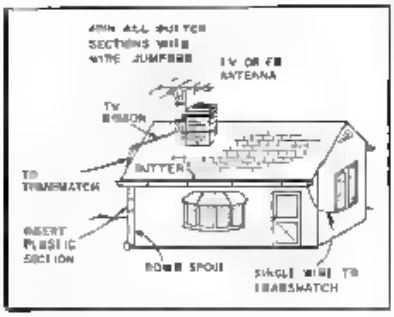

Fig. 3 — Some amateurs have had success when using the rain gutter as a random-length, end-fed radiator. The lower portion should be insulated from the remainder of the gutter and downspout to prevent rf hazards to animals and people. The TV or fm antenna-feed wire also can be used as a random-length antenna, as shown here.

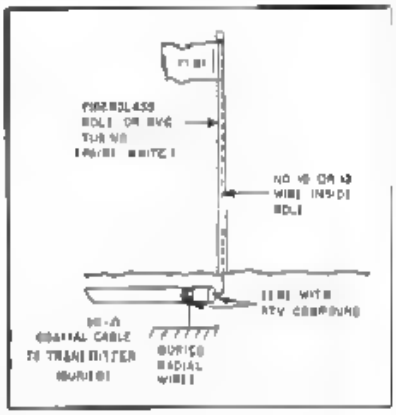

Fig. 4 — One can be patriotic while using a disguised antenna. The flagpole also serves as a vertical antenna in this illustration. Insulation should be used over the lower part of the antenna to prevent shock hazard to people and animals.

have to be on TV standoff insulators and spaced well away from phone and power company service-entrance lines. The TV or fm radio must of course be disconnected from the system when it is used for amateur work. Similarly, masthead amplifiers and splitters must be removed from the line if the system is to be used for amateur operation.

Flagpole Antenna

We can exhibit our patriotism and have an invisible amateur antenna at the same time by disguising our radiator as shown in Fig. 4. The vertical antenna is a wire that has been placed inside a plastic or fiberglass pole.

As shown, the flagpole antenna is structured for a single amateur band, and it is assumed that the height of the pole corresponds to a quarter wavelength for the chosen band. The radials and feed line can be buried in the ground as shown. In a practical installation, the sealed end of the

coax cable would protrude slightly into the lower end of the plastic pole.

If a large-diameter, fiberglass pole were available, we might be able to conceal a four-band trap vertical inside it. Alternatively, we might use a metal pole and bury at its base a water-tight box that contained fixed-tuned matching networks for the bands of interest. The networks could be selected remotely by means of a stepping relay inside the box. A 30-ft flagpole would provide good results in this kind of system, provided it was used with a buried radial system. At least one commercial antenna (from Delta Corp.) is used in this manner, but with an elaborate, continuously adjustable matching network (and VSWR indicator) that is operated remotely.

Still another technique uses a wooden flagpole. A small-diameter wire can be stapled to the pole and routed underground to the coax feeder or the matching box. The halyard could by itself constitute the antenna wire if it were made from heavy-duty insulated hookup wire. There are countless variations for this type of antenna, and they are limited only by the imagination of the amateur.

Other Invisible Antennas

Some amateurs have used the metal fence on apartment verandas as antennas, and have had good results on the higher hf bands (20, 15 and 10 meters). We must presume that the fences were not connected to the steel framework of the building, but rather were insulated by the concrete floor to which they were affixed. These veranda fences have also been used effectively as ground systems (counterpoises) for hf-band vertical antennas that were put in place temporarily after darkness fell.

One New York City amateur used the fire escape on his apartment building as a 40-meter antenna, and reported high success in working DX stations with it. Another apartment dweller made use of the aluminum frame on his living-room picture window as an antenna for 10 and 15 meters. He worked it against the metal conductors of the baseboard heater in the same room.

There have been many jokes told over the past decades about "bed spring antennas." The idea is by no means absurd. Bed springs and metal end boards have been used to advantage by many apartment dwellers as 20-, 15- and 10-meter radiators. A counterpoise ground can be routed along the baseboard of the bedroom and used in combination with the bed spring. It is important to remember that any independent (insulated) metal object of reasonable size can serve as an antenna if the transmitter can be matched to it. An amateur in Detroit, Michigan, once used his Shopsmith craft machine (about 5 feet tall) as a 10-meter antenna. He worked a number of DX stations with it when band conditions were good.

A number of operators have used metal curtain rods and window screens for vhf work, and found them to be quite acceptable for local communications. You'll have best results with any of these makeshift antennas when the "antennas" are kept well away from house wiring and other conductive objects.

From June 1983 QST, p 15:

Lightweight Trap Antennas — Some Thoughts

Portable multiband antennas need not be heavy and bulky. Small traps and light-gauge wire can provide a trap dipole that fits in a lunch bag. Try these practical guidelines for your next small antenna.

By Doug DeMaw, W1FB
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Vacationers, campers, sales people and QRPers take note! You need not carry a large multiband trap dipole afield if your transmitter is in the 150-W-output class, or lower. You can construct your own traps inexpensively with ordinary materials, and they can be made quite small without becoming poor performers. This article describes some easy techniques for fabricating homemade antenna traps. Additional hints are offered for keeping the bulk and weight of portable antennas within reason.

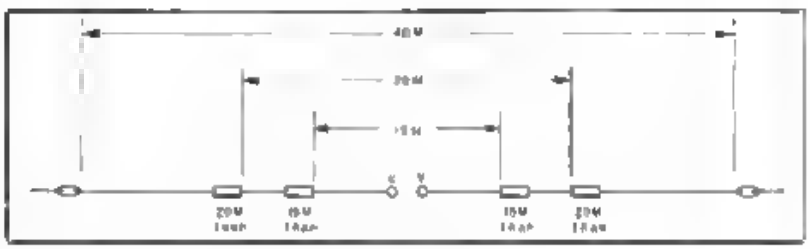

Fig 1 — Representation of a three-band trap dipole antenna

A Review of the Trap Concept

A "trap" is exactly what the term implies. It traps an rf signal to prevent it from passing beyond a specific point along an electrical conductor. At some other frequency, however, it no longer acts as a trap, and permits the passage of rf energy.

An antenna trap is designed for a particular operating frequency, and there may be several traps in the overall system — each designed for a specific frequency. Therefore, a 40- through 10-meter trap dipole might contain traps for 10, 15, 20 and 30 meters. On 40 meters, all of the traps are "absorbed" into the system to become part of the overall 40-meter dipole. Owing to the loading effect of the traps, the 40-meter portion of the antenna will be somewhat shorter than a full-size 40-meter dipole with no traps. The antenna bandwidth will be narrower when traps are used. Fig 1 illustrates the general format for a multiband dipole.

A trap style of antenna is not as efficient as a full size dipole. This is because there will always be some losses in the traps. But the losses in a well-designed system are usually so low that they are hard to measure by simple means. The losses represent a small tradeoff for the convenience of being able to accommodate many ham bands with one radiator and a single feed line. Yagi antennas contain traps in the parasitic elements (directors and reflectors) as well as in the driven element. Therefore, a multielement antenna of that type may have as many as 12 traps.

Electrical Characteristics

An antenna trap is a parallel-resonant L-C circuit. Therefore, it is similar to the tuned circuit in a transmitter or receiver. A resonator of this kind, if designed correctly, has a moderate Q and a fairly narrow bandwidth. This means that the trap capacitor should have a high Q and the trap coil should contain wire that is reasonably large in cross section. These traits will help to reduce losses.

Fig. 2 shows the equivalent circuit for an antenna trap. Once this network is adjusted to resonance in the desired part of an amateur band, it will not be affected significantly by the attachment of the wires that comprise the antenna. A well-designed and -constructed trap should not change frequency by any great amount when the temperature or humidity around it varies. Therefore, it is important to use a stable capacitor, a rigid coil and some type of sealant.

Mini Trap Using a Toroid Core

In an effort to scale down the size of my antenna traps during a design exercise for a portable antenna, I decided to investigate the worth of small toroid cores upon which to wind the coils. Ferrite cores were ruled out because they aren't as stable as powdered-iron ones. Furthermore, the powdered-iron material has a much greater flux density than an equivalent-size ferrite core, which means that the core will not saturate as easily at moderate rf power levels.

Development work started with Micrometals Corporation T50-6 toroids, which are sold by Amidon Associates, Palomar Engineers and RadioKit (see *QST* advertisements). My first effort resulted in a pair of very small 20-meter traps. A silver-mica capacitor was chosen for the parallel-tuned circuit. Ceramic capacitors were not used because of previous experiences I had with changes in value under temperature extremes; I had better results with dipped silver-mica units.

My rule of thumb for choosing the coil and capacitor values for traps is based on

a reactance of approximately 200 ohms, although values up to 300 have also yielded good results. Using 200 ohms as the basis for the design, I calculated the capacitor to be a value that was very close to a standard one — 56 pF for trap resonance at 14.100 MHz. This was obtained from

$$C(\mu F) = \frac{1}{2\pi f(MHz)\, X_C} \qquad \text{[Eq. 1]}$$

Hence

$$C = \frac{1}{6.28 \times 14.1 \times 200}$$

$$= 0.0000564\ \mu F\ (56\ pF)$$

Since X_C and X_L are equal at resonance, the coil was calculated by means of Eq. 2:

$$L(\mu H) = \frac{X_L}{2\pi f(MHz)} \qquad \text{[Eq. 2]}$$

Hence

$$L = \frac{200}{6.28 \times 14.1} = 2.25\ \mu H$$

(approximate)

The value of the coil will have to be adjusted slightly after the trap is assembled to allow for capacitor tolerance and stray capacitance, which accounts for the term "approximate" in Eq. 2.

The Amidon toroid tables were consulted to learn the A_L factor of a T50-6 core (1/2-inch-diameter toroid). The value is 40. From this I calculated the number of turns from

$$\text{Turns} = 100\sqrt{L_{\mu H}/A_L} \qquad \text{[Eq. 3]}$$

Hence

$$\text{Turns} = 100\sqrt{2.25/40} = 23.7$$

For practical reasons a 24-turn winding was used. A partial turn is not convenient on a toroid form.

The same procedure was used for the remaining traps in my antenna. This article is not a course in basic math, but the equations can be useful to those who have not previously designed resonant circuits or used toroidal cores.

Toroidal-Trap Adjustment

It's best to use the largest size wire that will fit easily on the toroid core. The stiffness of the heavier magnet wire will help to keep the coil turns in place, thereby minimizing detuning. I used no. 24 enameled wire.

The capacitor leads and coil "pigtails" should be kept as short as possible. Fig. 3 illustrates the layout I used. The leads at each end of the mica capacitor are soldered to the related coil leads before final adjustment is made.

A dip meter can be used to determine the resonant frequency of the trap, as shown in Fig. 4. Although a prominent feature of a toroidal coil is the self-shielding characteristic, which makes it difficult for us to get ample coupling with a dip meter, it is possible to read a dip. I have found that by inserting the dip-meter coil into the area of the winding gap on the tuned circuit (Fig. 4) a dip can be obtained. By approaching the trap from different angles it should be easy to find a spot where a dip can be read on the meter. Once the dip is found, back off the instrument until the dip is barely discernible (the minimum coupling point). Monitor the dip-meter signal on a calibrated receiver to learn the resonant frequency of the trap.

Select a part of the related amateur band for trap resonance. I adjust my traps for the center of the frequency spread I am most interested in. For example, I set my 20-meter traps for resonance at 14.025 MHz because I work only cw from 14.000 to 14.050 MHz. For phone-band coverage, I'd pick 14.275 MHz as the trap frequency. A compromise frequency for phone and cw operation would be 14.100 MHz. Owing to the trap Q, coverage of an entire band is not possible without having an SWR of 2:1 or greater at the band-edge extremes. The absolute bandwidth will depend on the trap Q and the Q of the antenna itself.

If the trap is not on the desired frequency, move the turns of the toroid coil farther apart to raise the frequency. Push them closer together to lower the frequency. An alternative method for finding the trap resonance is shown in Fig. 5. The trap being tested is connected to terminals x and y. The coupling is very light in order to prevent the test-circuit capacitance from appearing in parallel with the trap. For this reason the coupling capacitors are only 2 pF. The station transmitter is adjusted for the lowest power output that will provide a reading on M1. The VFO is then swept manually across the band. When the resonant frequency of the trap is located, the meter (M1) will deflect upward sharply, indicating resonance. Adjust the trap for a frequency that is approximately 5% lower than the desired one. This will compensate for the shunt capacitance presented by the 2-pF coupling capacitors.

When the coil turns are set in the correct manner, spread a bead of fast-drying epoxy cement across the turns on the two flat sides of the toroid. This will prevent unwanted position changes that could cause a shift in resonance later on from handling.

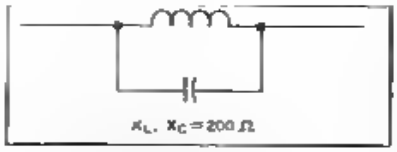

Fig. 2 — Electrical equivalent of an antenna trap. The ac resistance is not shown. A suitable reactance value for the coil and capacitor is 200 ohms.

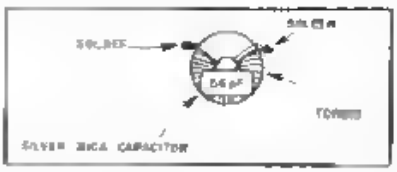

Fig. 3 — Physical arrangement for one of the toroidal L-C traps. Put spaghetti tubing over the capacitor leads to prevent them from shorting to the turns on the toroid.

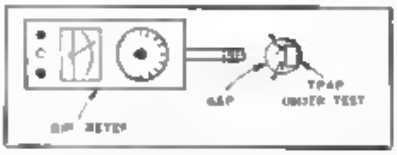

Fig. 4 — Test method for finding the resonant frequency of a trap. Different points around the toroid will yield better dip indications. Experiment with the position of the dipper coil.

Housing the Mini Trap

I learned that a 7/8-inch-OD PVC plumbing coupling, 1-1/4 inches long, would serve nicely as a housing for the toroidal traps.[1] A ridge inside the couplings at the center can be filed out easily to provide clearance for the trap. A rat-tail file does the job quickly. Fig. 6 shows a breakaway view of how the trap is assembled. Sheet of dowel rod are used for end plugs. A knot is tied in the antenna wire that enters the trap housing; this prevents strain on the trap coil.

After the antenna wire has been soldered to the trap at each end, add a layer of epoxy glue to the outer perimeter of one of the dowel plugs, then insert it into the PVC coupling until it is flush. Fill the coupling with nonconductive sealant; I used aquarium cement. Finally, place epoxy glue on the remaining end plug and insert it in the PVC coupling. Allow the trap to set for 48 hours, until the sealant has hardened. Fig. 7 is a photograph of a mini trap, along with a dipole center insulator made from a PVC T-coupling. The coupling is filled with sealant after the wires are soldered to the coaxial feed line. Long plugs are used to close the three open ends of the T connector. A closed nylon loop, made from strong spaghetti tubing, was fed through two small holes at the top of the T-coupling to permit erecting the dipole as an inverted V. A small eye bolt and nut could have been used instead.

There was a minor downward shift in trap resonance after the sealant hardened. Both 20-meter traps shifted roughly 30 kHz lower. No doubt this was caused by increased distributed capacitance across the coil turns with the sealant in place.

[1]Notes appear at end of article.

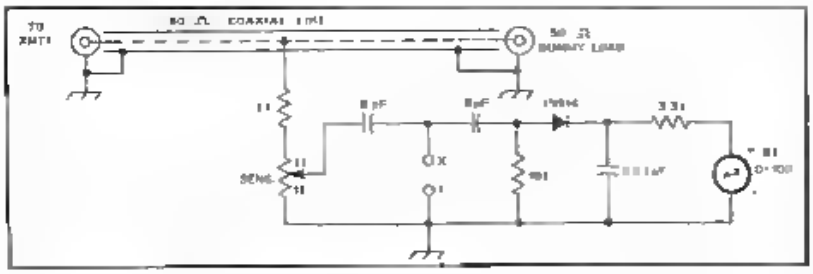

Fig 5 — Test fixture suitable for checking trap resonance with the station transmitter. Use the least amount of power necessary for meter deflection.

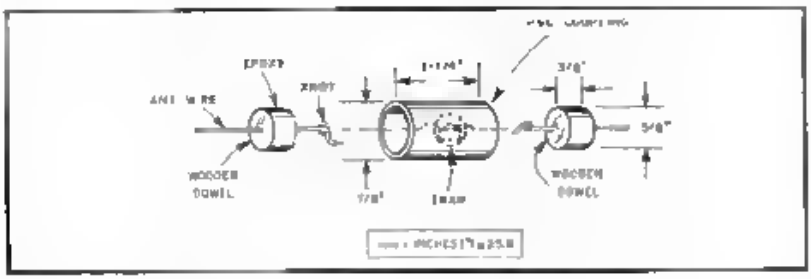

Fig 8 — Breakaway view of a toroidal mini trap. The knots in the wire prevent stress on the tuned circuit.

This seemed to have no effect on the trap quality; it had a measured parallel resistance of 25 kΩ before and after encapsulation (using the laboratory RX meter for tests). Generally, anything greater than 10 kΩ is suitable for an antenna trap.

Mini Coaxial-Cable Traps

Two very interesting articles concerning antenna traps appeared in the amateur literature during 1981.[2,3] After reading them a second time, I decided to attempt building some traps along the lines discussed in those articles. Some advantages over the usual coil/capacitor style of trap were described by the authors: (1) The traps were not especially frequency sensitive to changes in temperature and climate; (2) the coaxial trap offers greater effective bandwidth; and (3) parallel resistance is quite high — on the order of 50 kΩ.

The articles under discussion contained practical information about the use of RG-58/U and RG-8/U cable for the trap coils. I wanted a small, lightweight trap, so elected to see what could be done with miniature cable — RG-174/U. A completed mini coaxial trap for 20 meters is shown in Fig. 8.

The principle of operation is covered well by O'Neil (note 2). Since this article deals with the practical aspects of traps, we won't delve into the electrical characteristics of the coaxial trap too deeply. However, a diagram showing how it is hooked up is offered in Fig. 9B. A length of coaxial line is wound on a coil form, and the inner conductor at one end is attached to the outer conductor at the opposite end. The distributed capacitance of the two conductors and the inductance of the coil combine to provide a resonant circuit. An acceptable Q results, and the trap can accommodate considerable rf voltage and current without being damaged. A parallel resistance of 50 kΩ was measured for the 20-meter trap of Fig. 8. The bandwidth at the 10 kΩ points was somewhat greater than with the toroidal trap.

Coaxial-Trap Assembly

I found 5/8-inch-OD PVC plumbing pipe to be an acceptable and low-cost material for the coaxial traps. End plugs made from 1/2-inch wooden dowel fit snugly inside the PVC pipe. The completed trap contains a length of bus wire inside it for connecting the braid and center conductor of the cable together, as discussed earlier. The ends of the bus wire and the coaxial cable ends are routed outside the PVC tubing through small holes, then soldered. Aquarium cement was again used, this time to seal the six small holes drilled in the tubing. Epoxy cement was applied to the sides of the wooden plugs before inserting them into the tubing. A layer of vinyl electrical tape can be wound over the coaxial coil if desired, although this should not be necessary. If weather protection is desired, a coating of exterior polyurethane varnish can be applied to the completed close-wound coil. This will keep the turns affixed in the desired position after final adjustment.

Tune-up is carried out in the same manner as prescribed for the toroidal traps, using a dip meter or the test fixture described in Fig. 5.

The length of the coaxial cable used will have to be determined experimentally. My 20-meter coaxial trap contains 15 close-wound turns of RG-174 cable (36 inches, 89 pF) to provide resonance at 14 100 MHz. Final adjustment was done by moving the three inter turns at one end until the desired frequency was noted. The coil form for the 20-meter trap is 2-1/2 inches long. The wooden end plugs are 3/8 inch thick. The inside of this trap is not filled with sealant, but it could be if desired. Avoiding the use of fillet will make the traps lighter in weight, thereby permitting the use of lighter-gauge wire for the antenna sections.

Trap Performance

Both styles of trap were subjected to rf power tests to determine whether they could handle the output of a typical 150-W class transceiver. A Bird wattmeter was connected between the trap and the transmitter. A 50-ohm dummy load was attached to the opposite end of the trap. Next, 40- and 80-meter rf energy was applied (in separate tests) gradually while observing the reflected power, which of course was not conducive to providing an SWR of 1:1 with the trap in the line. Neither trap showed signs of heating or breakdown at power levels up to 150 W. A key-down period of five minutes was used during the tests, using a linear amplifier adjusted for 150 W output. Still no sign of power limitation. The SWR did not change under these conditions. I did not advance the power beyond 150 W, but it's safe to conclude that the coaxial-cable trap could sustain substantially more power without damage. This may not be true of the toroidal trap. I lacked the courage to find out!

Toward a Lightweight Dipole

Having solved the problem of lightweight, small traps I set about the task of reducing the bulk of the remainder of my multiband dipole. I am a dedicated miser, so the cost of materials was an important factor in the selection of wire and end insulators. I recalled a type of wire I had used on a number of DXpeditions: It was strong and light in weight, and the price was right! This wire is available from Radio Shack and similar outlets for use as speaker cable. It has a clear plastic outer covering, contains a no. 22 conductor (two each) and costs less than $5 per 100 feet. Hence, for this price we end up with 200 feet of wire (less than 2.5 cents per foot); the parallel conductors can be pulled apart easily without harming the outer insulation. In addition to the insulation aiding the strength of the wire portions of the antenna, it protects the copper

Fig 7 — View of a toroidal mini trap, an encapsulated toroid and a PVC T-coupling for use as a coaxial insulator. RG-58A/U cable is shown in this example (see text).

Fig 8 — A completed 80-meter coaxial trap with miniature RG-174/U coaxial cable.

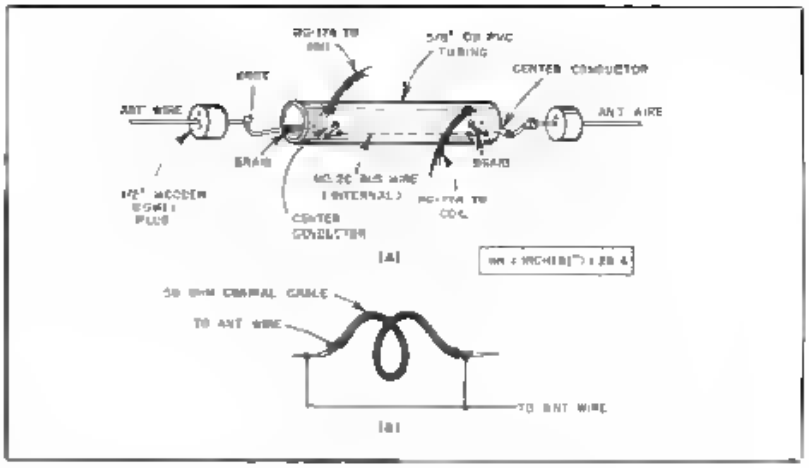

Fig 9 — At A is a breakaway view of a coaxial trap. The illustration at B shows the electrical connections for a coaxial trap.

wire from corrosion. This can be especially beneficial in areas where salt water and industrial pollutants affect the atmosphere. The Radio Shack number for this wire is 278-1385. I have observed no apparent deterioration of this type of conductor, even though some of my antennas have been aloft for three years.

Although RG-58/U coaxial cable is less offensive in terms of loss per 100 feet than is true of RG-174/U, we may want to trade losses for portability by using 174. Normally, a 50-foot length of feeder cable is adequate for portable work. In an effort to determine exactly what the hf-band losses per 50 feet might be, I tested this cable from 3.5 through 29 MHz. A Bird wattmeter was connected to each end of the 50-foot test cable. One wattmeter was terminated with a 50-ohm dummy load, and the other wattmeter was connected to a transmitter. The loss in decibels was as follows: 3.5 MHz — 1.19; 7.0 MHz — 1.42; 14.0 MHz — 1.67; 21 MHz — 1.93; 29 MHz — 2.0. Therefore, in a worst-case situation (10 meters), a 100-W power input to the cable would result in an antenna feed-point power of 63 W. RG-58/U, on the other hand, would have a 1-dB loss at 29 MHz, which would mean an antenna feed-point power of 79.4 W. This is not too significant when operating in the 50-150 W range, but it can be important when using a QRP rig with only a few watts or milliwatts of output power. I must say in defense of RG-174/U cable that I operated 20-meter cw with 2 W of output power from 8P6EU while using a dipole with 50 feet of RG-174/U feed line, and I worked the world without difficulty. I received many RST 599 signal reports. The tiny feeder cable and the hookup-wire dipole could be rolled up and stuffed in my hip pocket! The end and center insulators for that antenna were also lightweight. I made them from scraps of pc board from which the copper had been removed. The end insulators for the trap dipole discussed in this article were fashioned from inch-long pieces of 5/8-inch-diameter PVC tubing through which holes were drilled to accommodate the dipole wires and nylon guy lines.

Summary Comments

The overall length of any dipole section in a trap type of antenna will be less than if the dipole were cut for a single band without traps. The exception is the first dipole section after the feed point (out to the first set of traps). The following percentages (approximate) were typical in a coaxial-trap dipole I built for use from 40 through 10 meters, compared to the length of a full dipole (100%) for each band: 10 meters — 100%; 15 meters — 92.4%; 20 meters — 88.6%; 40 meters — 83.6%. The shortening becomes more pronounced as the frequency is lowered, owing to the cumulative loading effects of the traps.

These percentages can be applied during initial structuring of the antenna. Starting with the highest band, the dipole sections for each frequency of interest are trimmed or lengthened for the lowest attainable SWR. After the exact dimensions are known, continuous lengths of wire can be used between the traps. This will add strength to the antenna by avoiding breaks in the speaker-wire insulation, if that type of conductor is used. The percentage reductions listed above are not necessarily applicable to antennas that use toroids or other coil/capacitor traps. The wire diameter and insulation may also affect the final dimensions of the dipole.

For long-term installations, I would suggest the use of some type of sealer (spar varnish or polyurethane) over the wooden end plugs of the traps. All trap holes need to be sealed securely to prevent moisture from building up inside them.

Miniature antenna traps and lightweight trap dipole antennas are practical and inexpensive to build. Try one during your next vacation or business trip.

Notes

[1]mm = in × 25.4; m = ft × 0.3048

[2]G. O'Neil, "Trapping the Mysteries of Trapped Antennas," *Ham Radio*, Oct 1981, p 10.

[3]R. Johns, "Coaxial Cable Antenna Traps," *QST*, May 1981, p 15.

From December 1988 QST, p 31:

A Portable Vertical-Antenna Mount

Need a temporary, good-performing antenna? The mounting technique described here makes for quick installation of a multiband vertical with a minimum of fuss!

By Guy Black, W4PSJ
1201 Woodalda Dr
Winchester, VA 22601

Field Day contestants, vacationers and tenants sometimes need temporary antennas. On the high-frequency bands, a hook of wire hung from one or more trees often seems to be the best that can be done. To get such an antenna set up, a weight with a rope tied to it is usually thrown into a tree, and the antenna is then pulled up with the rope. I don't have very good aim and my throwing skills are underwhelming, so the dipoles and end-fed lengths of random wire I've managed to put up as temporary antennas have usually been disappointing performers.

Fig 1—Photograph of the portable vertical-antenna mount. The lower rotator-mast-mount casting is bolted to a steel plate, with a copper disc in between for attachment of radials. One of the bags of garden stone used to hold down the wooden stakes is visible in the upper right corner of the photo.

For the last several years, I have used a multiband vertical (a Butternut HF2V) at home with great success, particularly for DXing on the 80, 75 and 40-meter bands. Why not turn such a vertical into a portable antenna? It's essentially pre-tuned, and there's no strain on the throwing arm! Light, uncomplicated and easily transported multiband verticals have many possibilities.

My antenna came with a 22-inch aluminum ground stake. With care, it is possible to drive this stake into the ground repeatedly and without damage by using a short (one foot or more) section of TV masting, which fits nicely over the base insulator as a driver. Unfortunately, doing this requires the availability of a small sledgehammer (or a large one, depending on the ground!).

Another approach is to use a *portable base* with the vertical. This takes up a bit more space in the car than just a ground stake, but at least you don't need to carry along a sledgehammer!

Materials

The parts for my portable base are a 1 × 1-foot metal plate (an old rack panel works fine), the lower mast-support casting from a rotator (with its hardware), a 7-inch-diameter disc of copper-clad PC board, five 6-foot-long 1 × 1-inch hardwood stakes from a garden supply store, a 1 × 12-inch strip of thin hobby brass, a chassis mount SO-239 connector and a 6-inch length of 1-inch ID plastic water pipe. A few nuts, bolts, spade lugs, washers, and 500 feet of no. 14 stranded copper wire round out the materials list.

Construction

Bolt four of the hardwood stakes to the steel plate in a pinwheel configuration, as shown in Fig 1. Drill 10 equally spaced holes (large enough to pass no. 6-32 screws) around the outside edge of the PC-board disc. Then, center the mast support on the PC board, and mark and drill the four mounting holes on the PC board. Mark and drill the four mounting holes on the metal plate.

Thread 10 binding-head machine screws into the PC-board disc from the bottom, and connect the mast support, PC board disc and metal plate using the mast-support casting hardware. The heads of the machine screws should be between the metal plate and the PC-board disc. There is enough flex in the PC board so that the board won't break when the mast-support hardware is tightened. Slip the section of plastic pipe over the vertical-antenna mount, insert the pipe and antenna mount in the rotator mast support, and tighten the mounting clamps.

The section of plastic water pipe is necessary when using this mounting arrangement with a Butternut vertical, because the minimum diameter the clamps will grasp exceeds the one-inch OD of Butternut's base insulator. Other antennas may not require the plastic pipe section.

Part of the one-inch-wide brass strap is

used to connect the ground side of the vertical to the copper disc. The remaining piece of brass is used for mounting the SO-239 connector. (If your vertical already has a coaxial feed-point connector, skip this paragraph.) This brass piece should be about 1 × 2¾ inches, with a 5/8-inch hole near one end, and four no. 32 holes for no. 4-40 mounting hardware around the 5/8-inch hole. This hole is for mounting the SO-239 feed-point connector. Put a 90° bend in the brass piece about ¾ inch from the end opposite the SO-239 mounting hole. Mount the connector using no. 4-40 hardware, and solder the ¾-inch section of the brass strip to the disc of PC-board material so that the SO-239 faces away from the antenna and clears the hardwood stake (see Fig 1). Connect the shortest practical length of no. 14 wire from the SO-239 to the feed-point connection of the vertical.

Assemble the vertical according to the manufacturer's instructions, and install it on the mount. I'm not sure how much wind force the portable vertical antenna system can handle, so when I use it I weight down the stabilizer stakes with bags of garden stone, one of which can be seen in the upper right corner of Fig 1. An easier (and lighter) solution is to guy the antenna with lightweight nylon rope. Bricks (for holding the ends of the radials) and bags of garden stone or sand are widely available, and cheap enough to discard when you're through with them.

The Radials

I use the portable vertical-antenna mount with ten 50-foot radials, each spaced 36° apart on the ground. By using flanged, solderless spade lugs (Waldom DS-4083) it is not necessary to remove the nuts on the machine screws to connect the radials. The outer ends of the radials are held down by bricks. (Bricks aren't needed to hold down the radial wires if rocks or some other suitable weights are available.)

The fifth hardwood stake is used in lining up the radials (it also serves as a spare mounting-plate stabilizer). Paint a mark on each radial wire, 9 feet 5 inches from the machine-screw connection point (nail polish works fine for this). At that distance from the casting, uniformly space the radials 36° apart by laying the spare six-foot stake between the painted marks on adjacent radial wires. This makes for a neat layout with a minimum of fuss. Installation is easier if the radial wires are coiled up from the brick end (so that the connection lug is on the outside of the coiled radial). Tape or wire ties can be used to keep the coiled radials from getting tangled.

Results

With the Butternut HF2V erected on the portable mount in my back yard, tune up went smoothly. The antenna has an SWR of less than 2:1 across the 40-meter band, and over the selected 30 kHz of 75 meters. Switching back and forth between the portable vertical and my permanently installed vertical (also an HF2V, but with a larger radial field) I found little difference in received signal strength. I had the same good results working DX on both antennas. A 100-foot square area is needed for the ground radials if the antenna is put up as described, but the radials can be bent or shortened if necessary.

This antenna is so easy to put up and take down that it can be erected for just a few days' use. For easy transportation, I use the antenna's original 66-inch-long shipping carton to carry the antenna and the stakes, and a two-gallon milk crate for the radials, base plate and hardware. All the components of a handy and effective portable antenna system are in two packages, ready to go, and my throwing arm doesn't even get a workout!

From December 1997 QST, p 25:

An Extended Double Zepp Antenna for 12 Meters

Got a little over 50 feet of horizontal space to spare for a 24-MHz skywire? This simple antenna will beat your half-wave dipole by about 3 dB—and you can phase two of them for even more gain and directivity.

By John J. Reh, K7KGP
510 Mt Defiance Cir SW
Issaquah, WA 98027

According to *The ARRL Antenna Book*, *Zepp*—short for Zeppelin—is a term long applied to just about any resonant antenna end-fed by a two-wire transmission line.[1] A bit further on in the *Antenna Book*, there's a discussion of the extended double Zepp (EDZ) antenna.[2] This interested me because I have always been intrigued by "old-fashioned" wire antennas—and because the old-fashioned extended double Zepp's 3-dB gain over a half-wave dipole would provide performance quite suitable for *modern* times! The EDZ antenna consists of two collinear 0.64-λ elements fed in phase. Fig 1 shows current distribution in an EDZ, and Fig 2 shows the EDZ's horizontal directivity pattern in free space.

The extended double Zepp's theoretical performance looked good to me, so I designed and built an EDZ antenna for the 12-meter band. Fig 3 shows its configuration. I decided to cut mine for 24.950 MHz. Each EDZ element is 25 feet, 3 inches long, and consists of no. 14 stranded copper wire. The antenna elements are center-fed by a short matching section made of a 5-foot, 5-inch length of 450-Ω open-wire line. Connection to 52-Ω coaxial feed line is made by means of a 1:1 balun transformer. My EDZ is strung between two trees, 35 feet above ground.

[1]Notes appear at end of article.

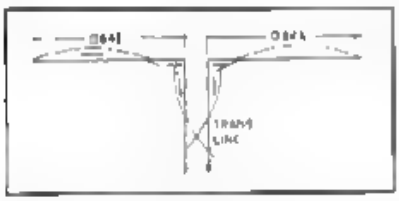

Fig 1—The extended double Zepp antenna consists of two 0.64-λ elements fed in phase.

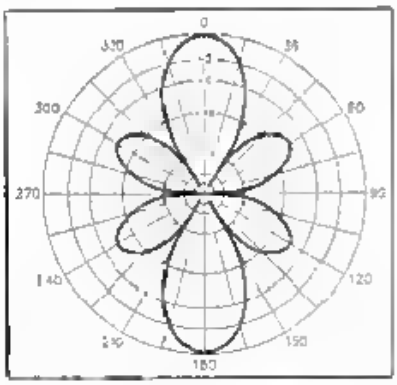

Fig 2—Horizontal directivity pattern for an extended double Zepp antenna in free space. Relative to a half-wave dipole, it exhibits a gain of approximately 3 dB. The antenna elements lie along the 90°-270° line.

Matching Section

Perhaps I am "reinventing the wheel," but I have not seen this matching method

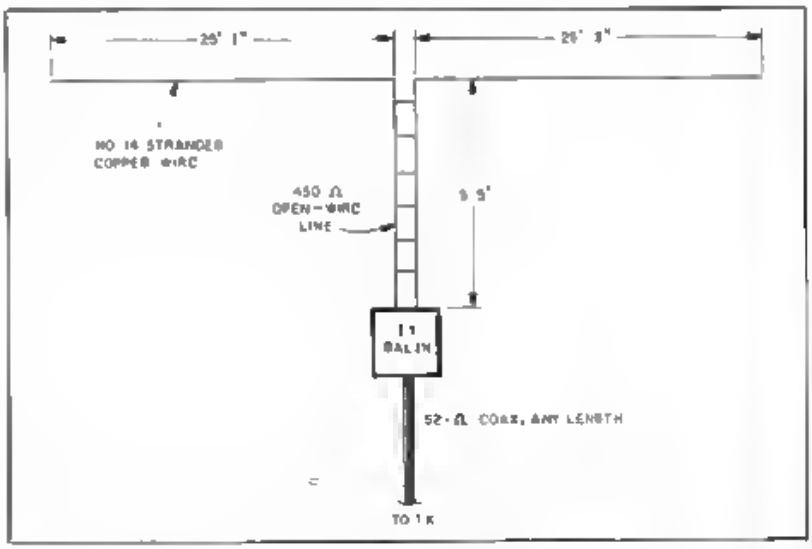

Fig 3—The extended double Zepp at K7KGP cut for 24.950 MHz. The 450-Ω matching section transforms the EDZ's calculated input impedance (142-j555 Ω) to 65 Ω (measured) for connection to 52-Ω coaxial cable by means of a 1:1 balun. The electrical length of the matching section is 52°; the linear dimension shown in the drawing assumes 450-Ω line with a velocity factor of 0.95.

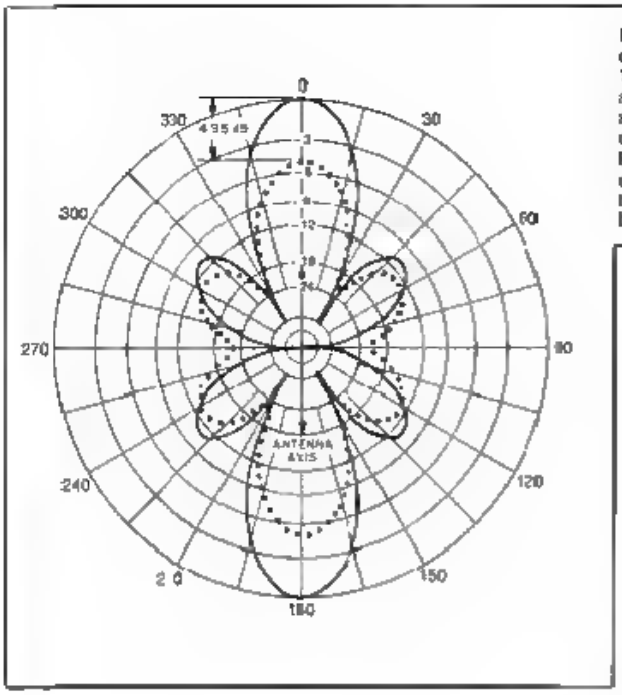

Fig 4—Comparison of calculated horizontal directivity patterns of one extended double Zepp (dotted line), and two EDZs spaced at 1/8 λ and fed 180° out of phase (solid line). The antenna axes lie along the 0°-180° line, and the antennas are mounted 35 feet above average earth. The phased EDZs exhibit nearly 5 dB gain over a single EDZ. This is 7 to 8 dB gain over a half wave dipole. Beamwidth of the two-EDZ array is 30°. The antenna axis is the same for the single EDZ and both EDZs in the phased array. The two-EDZ configuration characterized here is an *end-fire* array because maximum radiation occurs along its axis.

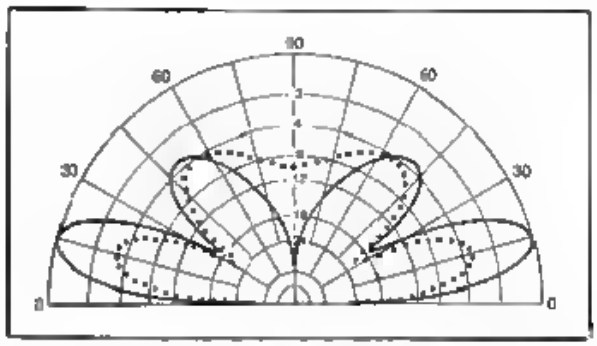

Fig 5—Comparison of the calculated vertical directivity patterns of one EDZ (dotted line), and two EDZs spaced at 1/8 λ and fed 180° out of phase (solid line). The antenna axis lies along the 0° line.

elsewhere.[3] The open-wire-line matching section is 52 electrical degrees long (0.145 λ). The matching section transforms the EDZ's input impedance to about 55 ohms, as measured with a noise bridge. The matching section dimension given in Fig 3 assumes a velocity factor of 0.95 for the 450-Ω line.

Trimming the matching section to size is the only adjustment necessary with the EDZ. Make the transformer a little long to begin with, and shorten it an inch or two at a time to bring the system into resonance. (You can check resonance with a noise bridge or by monitoring the SWR.) Do *not* change the length of the elements—the EDZ's gain and directivity depend on its elements being 0.64 λ long.

Phasing Two EDZs for More Gain and Directivity

Properly phased, two extended double Zepp antennas can give improved gain and directivity over a single EDZ. Fig 4 compares the calculated horizontal directivity patterns of a single EDZ and an array consisting of two EDZs spaced at 1/8 λ and fed 180° out of phase. Fig 5 compares the vertical radiation patterns of the single and phased EDZs.

Fig 6 shows the dimensions of a practical two-EDZ configuration. With proper adjustment, it exhibits an SWR of 1.3:1 across the 24-MHz band. In the array I built, lightweight broom handles serve as spreaders between the element ends, the center spreader is a wooden slat. I used nylon rope to haul the array up between two trees. This antenna system works well, but poor propagation has precluded a thorough tryout so far. The contacts I have had with it have been entirely satisfactory.

The matching method shown in Fig 6 is somewhat clumsy because the combined length of the phasing lines is greater than the spacing between the EDZs. The feed method shown in Fig 7 should be easier to build because the combined length of the phasing lines equals the spacing between

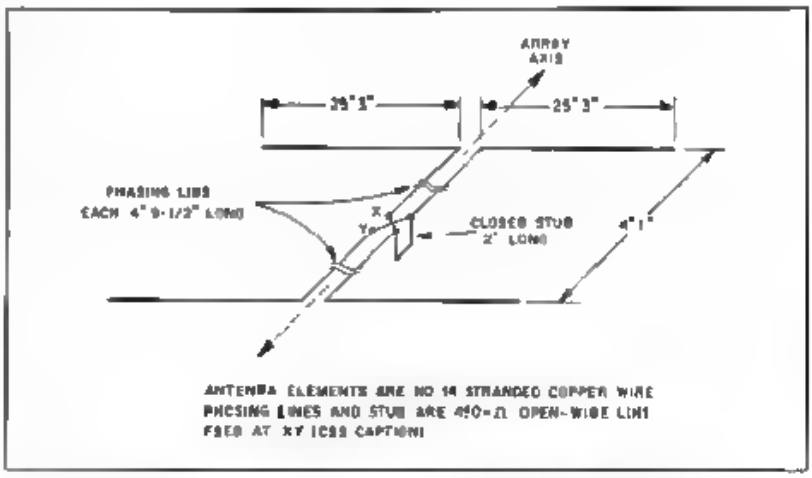

Fig 6—One method of phasing two EDZs for greater gain and directivity. The array is bidirectional, with maximum radiation occurring along the array axis. The impedance across points X and Y is 50 Ω, balanced; with a 1:1 balun at XY, the array can be fed by means of 52-Ω coaxial cable. The stub, 1.5° long, cancels a capacitive reactance of approximately 13.5 Ω at the feed point. This array works well, but its matching system is clumsy because the combined length of the phasing lines is greater than the spacing of the two EDZs. Fig 7 shows a proposed feed method that takes up less space.

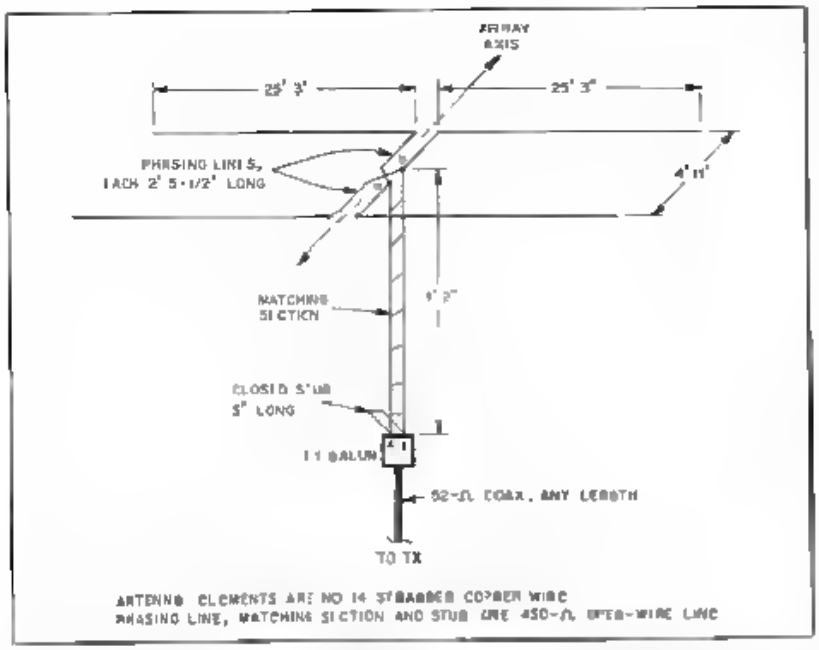

Fig 7—Proposed alternative method of phasing two EDZs. In this arrangement, the length of each phasing line is half the EDZ spacing. Calculated impedance across points a and b is 15-j112 Ω. The matching section—11' in length—transforms this to a calculated impedance of approximately 55-j32 Ω (balanced) across points x and y. The stub, 4" long, cancels the capacitive reactance (32 Ω). A 1:1 balun transformer allows the array to be fed by means of 52-Ω coaxial cable. See text.

the EDZs. I have not tried this matching method, but I'm confident that my calculated dimensions are close to what will actually be encountered.[4]

Conclusion

If the extended double Zepp has caught your attention, but 12 meters hasn't, you can scale the linear dimensions given here for other bands of interest. Once your EDZ is up and working, I think you'll agree that the performance of the "old-fashioned" extended double Zepp isn't old-fashioned at all!

Notes

[1]*The ARRL Antenna Book*, J. Hall, ed. (Newington, ARRL, 1984) p 5-4.

[2]*The ARRL Antenna Book*, p 6-5.

[3]K7KGP's matching technique is a "reinvention of the wheel" of which he can be proud. Termed the series section transformer, it appears in *The ARRL Antenna Book* and *The ARRL Handbook*. The series-section material in these books is based on Frank A. Regier, "Series-Section Transmission-Line Impedance Matching," *QST*, Jul 1979, pp 14-16.—Ed.

[4]K7KGP's calculations were confirmed by Rus Healy, NJ2L, of the ARRL HQ Technical staff, using the Smith® Chart and the Mini-Numerical Electronics Code (MININEC) on an IBM® personal computer. Data for the plots in Figs 4 and 5 were also generated by means of MININEC.—Ed.

From December 1988 QST, p 47:

Scaling the Extended Double Zepp

☐ You can easily scale the design of an extended double Zepp (EDZ)[1] to work on another band. For example, assume you wanted to build an EDZ for 7.2 MHz, basing the design on the 24.95-MHz antenna presented in my December article. The 24.95-MHz antenna has element lengths of 25' 3" and the matching-transformer line length is 5' 5". Use the following formula to scale the antenna dimentions to the desired band:

$$L2 = \frac{(f1 \times L1)}{f2} \quad \text{(Eq 1)}$$

where

L2 = length at the desired frequency

f1 = resonant frequency of the original antenna

L1 = length of interest at the resonant frequency of the original antenna

f2 = resonant frequency of the new antenna

Lengths for L1 and L2 must be expressed in similar units of measurement (feet, meters, etc); this also applies to the units of measurement used for f1 and f2 (kilohertz, megahertz, etc).

Substituting the values for element length

$$L2 = \frac{24.95 \times 25.25}{7.2} = 87'6'' \quad \text{(Eq 2)}$$

and matching-transformer length

$$L2 = \frac{24.95 \times 5.417}{7.2} = 18'9'' \quad \text{(Eq 3)}$$

This scaling technique also works for element spacings. Velocity-factor considerations can be ignored because they were included in the initial design.—*John Reh, K7KGP, 510 Mt Defiance Cir SW, Issaquah, WA 98027*

[1]J. Reh, "An Extended Double Zepp Antenna for 12 Meters," *QST*, Dec 1987, pp 25-27.

From October 1988 QST, p 42

An Indoor Dipole Antenna

☐ I live in an apartment. Because of this, I'm limited in the size and type of antenna I can install for use on HF. After trying end-fed random wires, loops, mobile verticals, rain gutters and so on, I designed a multiband dipole antenna that requires no tuning after installation. It's inconspicuous, non-hazardous and efficient. I used the following materials to construct it: one PL-259 connector; 12 feet of "Mini 8" coaxial cable; two nylon cable ties, approximately 45 feet of no. 22 insulated, solid copper wire; six test leads with alligator clips; 26 thumbtacks; and an SWR bridge. The antenna was installed in less than two hours.

After attaching the PL-259 to the coaxial cable, I wound 6 feet of the coax into a tight coil and held this winding together with two nylon cable ties. The result is a shield-choke balun at the point where the antenna elements attach to the cable.[2]

Using the formula f (feet) = 234 ÷ f (MHz), I calculated the length of wire necessary for each leg of a half-wave dipole at 21.1 MHz. Next, I cut two wires to this length and attached them to the feed line, one to the shield braid and the other to the center conductor. Using my transmitter and SWR meter, I pruned the dipole ends equally until I obtained the lowest possible SWR at 21 MHz. (Caution: Trim the antenna wires only when the transmitter is *off*.)

At this point, the clip leads come into play. To get the antenna up and running on 14 MHz, follow this procedure. (1) Attach a clip lead to the end of the 15-meter dipole; (2) calculate the length of the legs of a 14-MHz dipole; (3) add enough wire to each clip lead/dipole leg to bring the total length of the each 14-MHz dipole leg to the length calculated in step 2; and (4) prune the added wire for maximum SWR at the 14-MHz design frequency with the aid of the transmitter and SWR bridge. Continue this procedure to add additional clip leads and wire segments for 10 and 7 MHz. I used the thumbtacks to secure the wire pieces and test leads to the plasterboard ceiling of my apartment. Fig 6 shows the configuration of the entire antenna in linear form.

In my installation, the actual length of the dipole legs for a given band is about 14% shorter than the calculated length. This is probably due to the proximity of the antenna to the apartment ceiling—and the fact that I had to install the antenna around the perimeter of a square room, almost like a loop!

Careful pruning of the antenna for my favorite band segments paid off: An antenna tuner is unnecessary on all of the antenna's four bands. With the addition of Doug DeMaw's "AC Outlet Strip with Filtering" (December 1986 *QST*, pages 25-27), I eliminated TVI and RFI from my station.

—Larry A. Barry, NV5I, 5903 Danny Kaye #1308, San Antonio, TX 78240

[2]See Bob Schetgen, "Shield Chokes for Coaxial Cable," *QST*, Mar 1988, p 41.

An antenna similar to Larry's has been in use at AK7M for several years. I use alligator clips instead of test leads and my antenna's wire sections are held away from the plasterboard by nylon cable ties and thumbtacks. I can't complain about its performance. I've worked plenty of DX on 30, 20 and 15 meters running just 20 W output. Moral: All's not lost if you live in an apartment. Just keep plugging away with That Old Ham Spirit!—*AK7M*

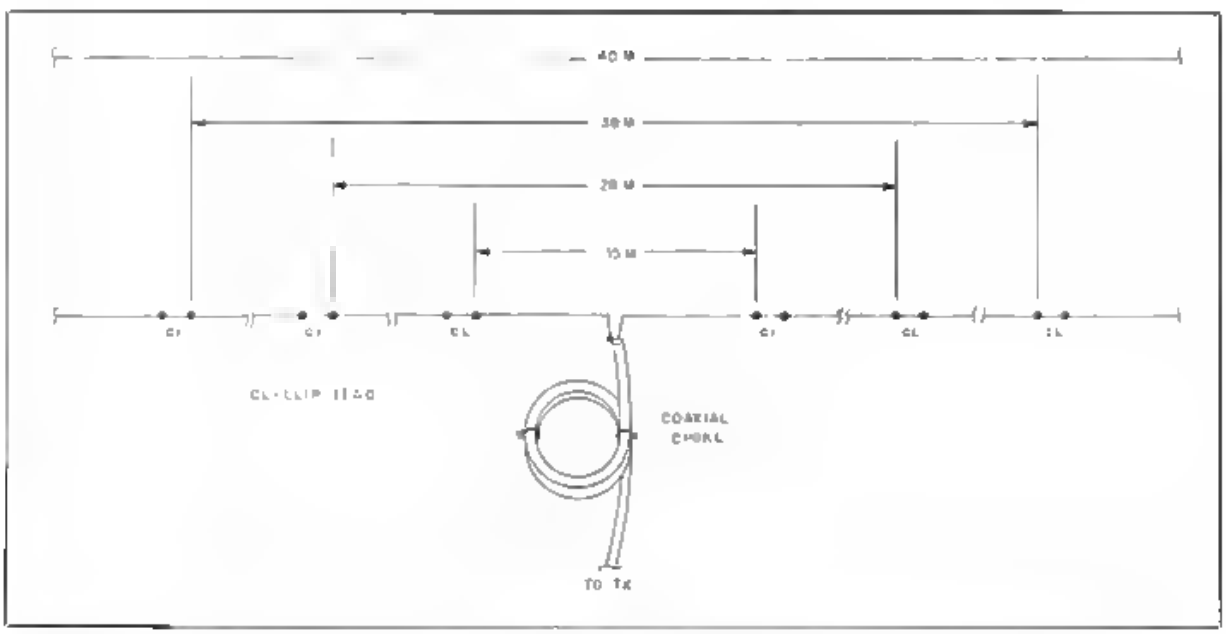

Fig 6—Larry Barry's multiband dipole makes crafty use of clip leads and thumbtacks to stuff half-wave dipoles for 15, 20, 30 and 40 meters into cramped apartment space. Changing bands entails only the connection or disconnection of clip leads. This drawing shows a straight dipole. Larry's antenna is bent into a square but works just fine. See text.

From April 1989 QST, p 38:

A Short 7-MHz Dipole

☐ Here are dimensions and construction information for a short, inductively loaded dipole for 40 meters. If installed over 50 ft above ground—outdoors or even in an apartment—it can provide plenty of DX.

See Fig 2. The antenna and loading coils consist of a total of 60 ft of no. 14 plastic-covered wire. Wind the loading coils first: Each consists of 30 close-wound turns on a 1½-inch-diam plastic form (pill bottles are suitable—*AK7M*). Use the rest of the wire as shown in Fig 2. (If space prohibits an overall antenna length of 32½ ft, you can let the 6¼-ft end sections dangle for a total length of just over 20 ft. Feed the antenna as close to its center as you can: 50- or 72-ohm coax is suitable. Preferably, the feed line should leave the antenna at a right angle.

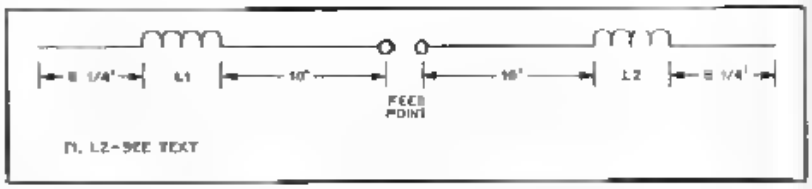

Fig 2—Stan Grimes suggests using this short, loaded 7-MHz half-way dipole where space is limited. The antenna and loading coils consist of no. 14 insulated wire, as in text.

This system can handle up to 120 W. Installed as shown in Fig 2, it should exhibit better than a 2:1 SWR from 7050 to 7160 kHz.—*Stan Grimes, W7CQB, 13300 NW 14th Ave #A, Vancouver, WA 98685-1652*

From July 1980 QST, p 17:

Active Filters

Why not build one of these nifty filters or use the design information to customize your own!

By Allan Bloom* N1AL

One of the triumphs of modern technology is that you can build "tuned circuits" and all kinds of other filters entirely without coils. Those generations of RTTY enthusiasts who grew up depending on the ubiquitous 88-mH toroidal inductors might be shocked to discover that you can replace up to four of these bulky items with a single IC. Besides their size and expense, coil-capacitor filters at audio frequencies are notoriously hard to tune — it's just hard to find variable coils or capacitors big enough to do the job. Many active filters can be tuned with an inexpensive potentiometer.

What *is* an active filter? Well, what is a *filter*? We generally consider a filter to be any circuit designed to eliminate some frequencies more than others. A *high-pass* filter passes high frequencies with little attenuation while providing great attenuation to the lower frequencies. See Fig. 1A. The *cutoff frequency* of a high-pass filter is the lowest frequency that passes with relatively little attenuation. The region above the cutoff frequency is the *passband* and the region of high attenuation is the *stopband*. A *low-pass* filter has its passband below the cutoff frequency and its stopband above. A *band-pass* filter has two stopbands — one above and one below the passband, and a *band-stop* filter has a stopband between a pair of passbands. See Fig. 1B.

An *active* filter is simply a filter that uses gain (active devices) to improve the attenuation characteristics. That Q-multiplier in your old receiver is an early type of active filter. While most active filters these days use operational amplifiers (op amps),[1] you can make some type of active filter with almost any device that has power gain.

RC Active Filters

It's quite possible to design active filters using coils. We've already mentioned the antidistortion Q-multiplier as one example, and we'll look at a couple of others later. But today, most people try to design inductors *out* of their circuits, at least at audio frequencies. As previously mentioned, coils for audio frequencies are often large and frequently expensive. Although passive LC (inductance-capacitance) filters require no power supply, you have to design them carefully to minimize loss, paying careful attention to input and output impedance matching. Active filters, on the other hand, can easily be designed for almost any desired input and output impedances, and can give considerable gain too. RC (resistance-capacitance) active filters are especially useful at low audio frequencies where the large inductances needed for LC filters become impractical.

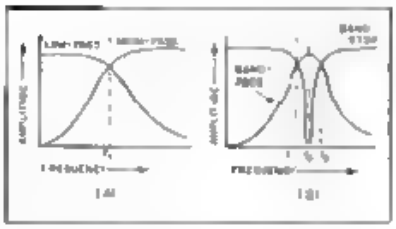

Fig. 1 — At A, plots of relative output versus frequency for high-pass and low-pass filters. f_c is the cutoff frequency. At B, plots of relative output versus frequency for band-pass and band-stop filters. f_0 is the center frequency. The area between f_1 and f_2 is the passband of the band-pass filter and the stopband of the band-stop filter.

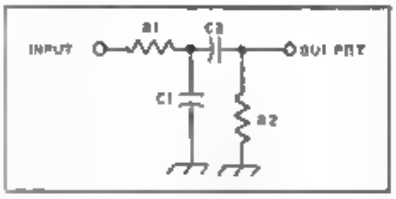

Fig. 2 — A passive RC band-pass filter. Maximum Q obtainable is only 1/2.

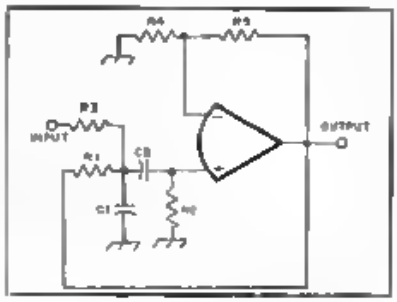

Fig. 3 — An active RC band-pass filter. To design a filter using this circuit, make all the frequency-determining resistors and capacitors equal: R1 = R2 = R3 and C1 = C2. Choose a convenient value for C and then $R = \sqrt{2}/2\pi f_0 C$, where R is in kΩ, C is in μF, and f_0 is in kHz. $Q = f_0/B$, where B is the 3-dB bandwidth in kHz. R4 and R5 determine the Q: R5/R4 = 3 − (√2/Q). The voltage gain is 2Q/√2 − 1.

Band-Pass RC Active Filters

You *can* make an RC filter without any active devices. Look at Fig. 2. At high frequencies, most of the signal is shunted out by C1. At low frequencies, most of the signal is blocked by C2. Thus the circuit of Fig. 2 is a band-pass filter. The limitation is that the maximum Q possible with this type of filter is only 1/2.

Those familiar with Q-multipliers or regenerative detectors may recall that one way to increase the Q of a tuned circuit is to introduce a little positive feedback around it. (If you apply too much feedback, the circuit will oscillate.) The same trick works for an RC bandpass filter. See Fig. 3. Here R3 has been added to couple in the signal. You can use a number of different resistor and capacitor values to achieve the desired filter characteristics, but for simplicity we usually make the two capacitor values the same and also let R1 = R2 = R3.

Let's say we want a 1-kHz band-pass filter with a 3-dB bandwidth of 600 Hz. The bandwidth is just the center frequency divided by the Q, so we have $B = f_0/Q$ or $Q = f_0/B = 1000/600 = 1.67$. So using the equations from Fig. 3, $R5/R4 = 3 - (\sqrt{2}/1.67) = 2.15$. The actual values of these resistors are not so important — it's the *ratio* of the two that determines

*1578 Los Alamos Rd, Santa Rosa, CA 95401

[1]Notes appear at end of article

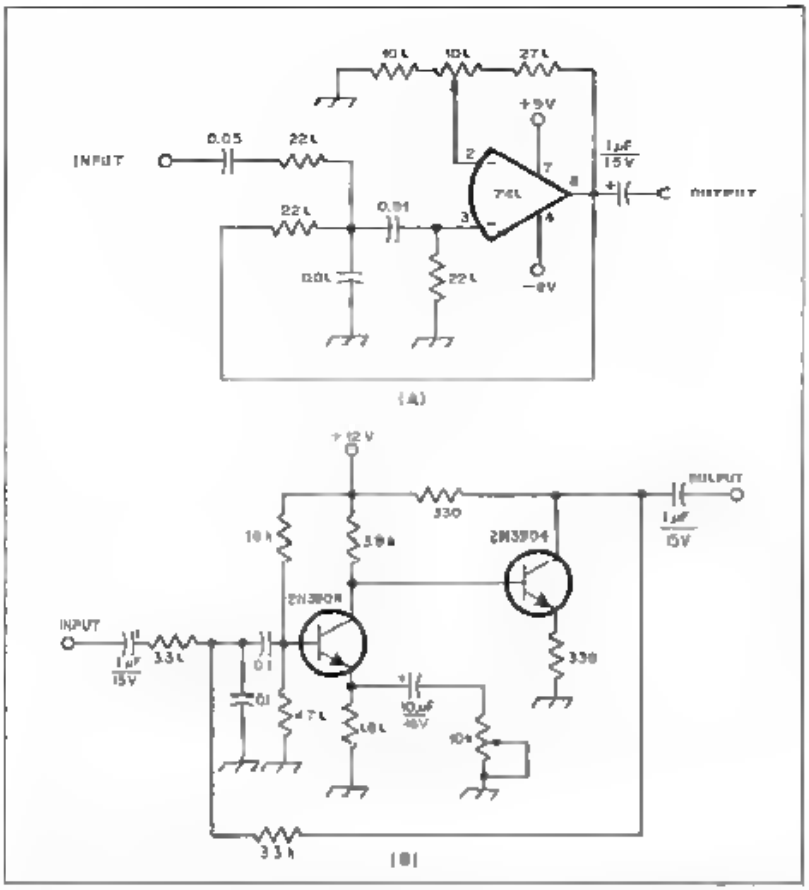

Fig 4 — A practical audio filter is shown at A, based on the design in Fig 3. The Q can be varied by adjusting the 10-kΩ potentiometer. A band-pass filter using discrete transistors is shown at B. R2 in Fig 3 is the parallel combination of the 4.7-kΩ and 10-kΩ resistors in Fig 4B (about 3 kΩ).

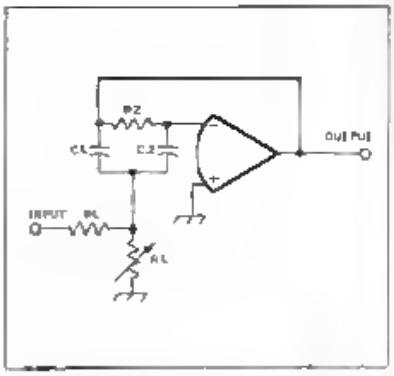

Fig 6 — A tunable band-pass filter. After choosing a value for C (C1 = C2), then R2 = 1/πBC, where R2 is in kΩ, C is in µF, and B is the bandwidth in kHz. R1 (kΩ) = R2/2G, where G is the desired numerical voltage gain at resonance.

$$R1\ (k\Omega) = \frac{1}{2\pi C\,(2f_0^2/B) - 8G}$$

where f_0 is in kHz. Insert the minimum and maximum values of f_0 into the above equation to get the maximum and minimum values for R1.

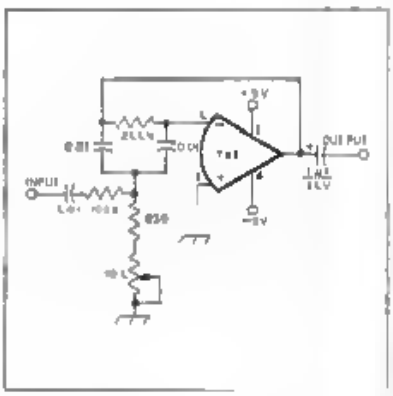

Fig 7 — A practical band-pass filter that tunes from 350 to 2000 Hz.

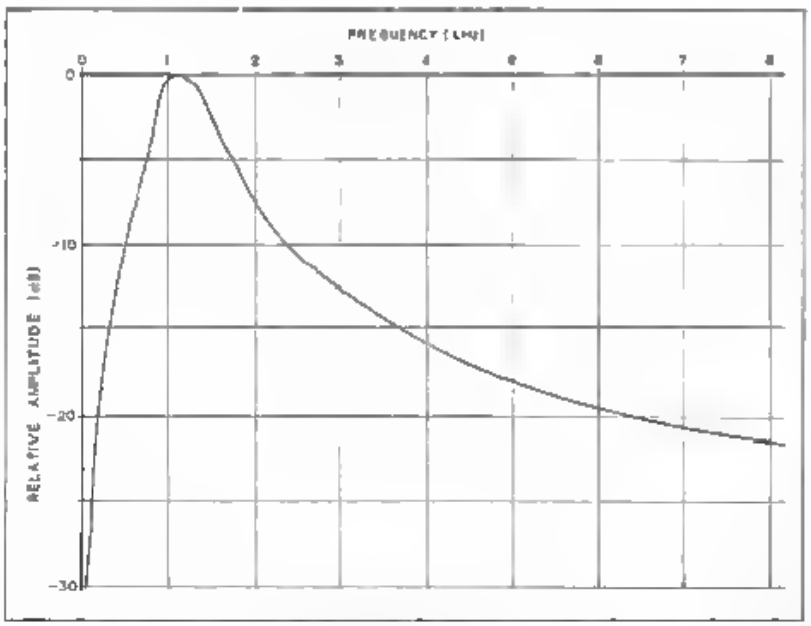

Fig 5 — Measured frequency response of the filter of Fig 4A. The center frequency and bandwidth are not exactly as predicted because of component tolerances.

the Q and gain. Let's choose R4 = 15 kΩ. Then R5 = 2.15 × 15 kΩ = 32 kΩ. (If R5/R4 = 3, the Q is infinite and the circuit becomes an oscillator.) To allow for resistor tolerances you usually use a potentiometer to adjust the gain to get the exact Q you want. See Fig 4A. With the potentiometer set to the middle of its range, the effective values of R4 and R5 are 15 kΩ and 32 kΩ respectively, as desired. Next, choose a value for R or C. Let's let C = 0.01 µF. (All of the formulas in this article express capacitance in microfarads, resistance in kilohms, and frequency in kilohertz.) Then R = $\sqrt{2}/(2\pi C f_0)$ = 22.5 kΩ or about 22 kΩ. Fig. 5 shows the measured frequency response of the circuit in Fig 4A. You can raise or lower the Q by adjusting the potentiometer. If you want to tune this

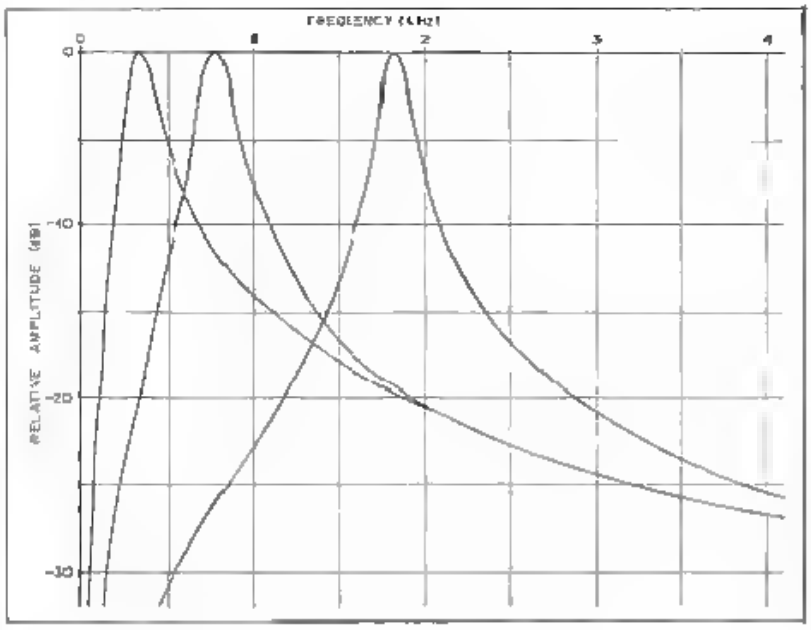

Fig. 8 — Measured frequency response of the circuit of Fig. 7 for three settings of the potentiometer.

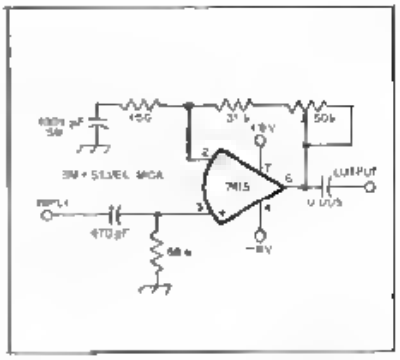

Fig. 9 — A 50 kHz band-pass filter. Calculated voltage gain is 200 and the Q is 10, giving a bandwidth of 5 kHz. Since a standard 741 op amp does not work well above 10 kHz, a high slew-rate version is used. To design for other frequencies, first choose a value for C1. Then R2 = 3B/2πC1f_0^2, where GB is the gain bandwidth product of the op amp (1000 kHz for a 741 or 741S). Choose Q using Q = f_0/B or Q = GB/f_0G. Then

$$R1 = \frac{R2}{GB}\left|\frac{1}{Q} - \frac{f_0}{GB}\right|$$

The highest possible Q is GB/f_0 and the highest possible gain is QGB/f_0.

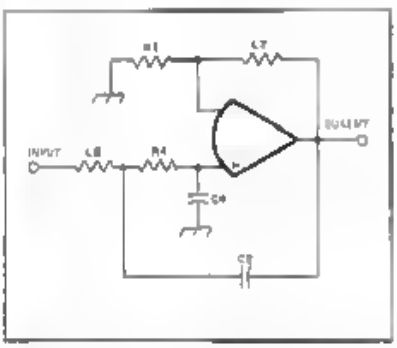

Fig. 9 — A low pass active filter. If Q > 1.0 (pick it then one of a few pass filter has a peak in the frequency response almost to that of a band-pass filter. For relatively narrow bandwidths Q is approximately f_0/B. R2/R1 = 2 − (1/Q). For a given value of C (C1 = C2), R3 = R4 = 1/2πf_0C), where R is in kΩ, C is in μF and f_0 is in kHz. The gain at f_0 is 3Q − 1.

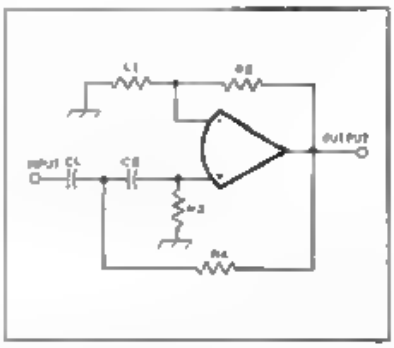

Fig. 11 — A high pass filter. For fairly narrow bandwidth (high Q), the Q of a high-pass filter is approximately f_0/B. R2/R1 = 2 − (1/Q). For a given value of C (C1 = C2), R3 = R4 = 1/2πf_0C), where all quantities are expressed in the same units used in the previous examples. The gain at f_0 is 3Q − 1.

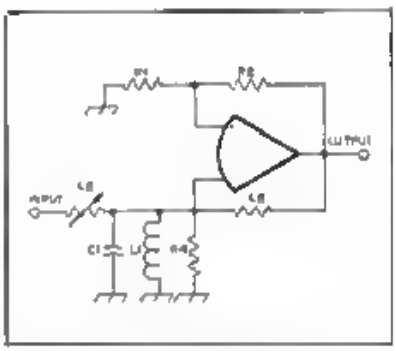

Fig. 12 — A band pass LC active filter. Q = f_0/B. Choose a convenient value for L1. Then

$$C1 = \frac{1}{(2\pi f_0)^2 L1}$$

where L1 is in henrys and C1 is in μF. R3 = QX, where X is the inductive or capacitive reactance in kΩ (X = 2πf_0L1). The gain is 1 + R2/R1. R2/R1 = R5/R4. R4 includes the losses in L1.

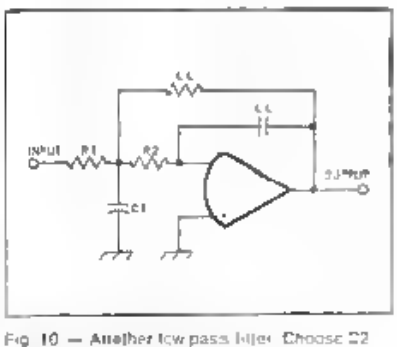

Fig. 10 — Another low pass filter. Choose C2 then C1 = C2/3Q^2. R1 = R2 = R3 = 1/(2πf_0 √(C1C2)), where R is in kΩ, C is in μF and f_0 is in kHz. The gain is equal to Q.

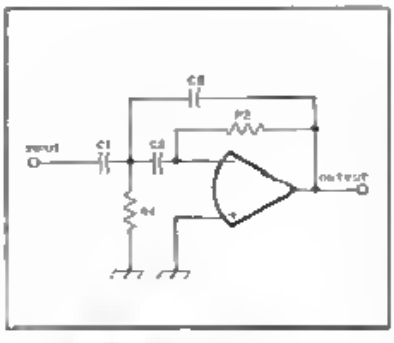

Fig. 12 — Another high-pass filter. Choose R1 then R2 = R1/3Q^2. C1 = C2 = C3 = 1/(2πf_0 √(R1R2)). The gain is equal to Q.

filter without changing the Q, you would need three ganged potentiometers to replace R1, R2 and R3.

Don't get the idea that all RC active filters must be made with op amps. The design of Fig. 3 works fine using a pair of transistors. Fig. 4B is a practical example. The center frequency is about 700 Hz, and the bandwidth is determined by the setting of the 10-kΩ potentiometer.

The filter of Fig. 6 has the interesting property that you can tune the center frequency without changing the gain by varying a single resistor, R3. In addition, the

Q increases with frequency in such a way that the bandwidth stays constant for all tuning settings — a sort of "poor man's passband tuning!"

To design one of these filters, you first choose the bandwidth (B), gain (G) and the lowest and highest frequencies to be tuned (f_{min}, f_{max}). Let's say you want to tune 350 to 2000 Hz (0.35 kHz to 2 kHz) with a bandwidth of 150 Hz (0.15 kHz) and a gain of one. Again we'll choose 0.01 μF for the capacitor value. From the formulas in Fig. 6, R2 = 1/(π × 0.15 × 0.01) = 212 Ω, R1 = 106 kΩ and the minimum and maximum values of R3 turn out to be 300 Ω and 10.7 kΩ. Using the nearest standard resistor values, we get the circuit of Fig. 7. Fig. 8 indicates the measured frequency response for the circuit. If your calculations give you a negative value for R3, then your lower frequency limit is too low or your gain is too high. Choose new values and recalculate.

Low-Pass RC Active Filters

If you need attenuation of *higher* frequencies only (such as adjacent-channel ssb interference), a low-pass filter will fill the bill. Representative designs are given in Figs. 9 and 10.

The circuit of Fig. 9 can be tuned by ganged potentiometers at R3 and R4. The Q can be adjusted by inserting a potentiometer between R1 and R2 as in Fig. 4A.

While it's not as easy to tune, the circuit of Fig. 10 has better stability than that of Fig. 9. For high values of Q, the gain and Q of the latter filter will change markedly for small changes in any of the resistor or capacitor values. If you need only a fixed-frequency filter, the one in Fig. 10 is a better choice.

High-Pass RC Active Filters

In principle, you can convert any RC low-pass filter into a high-pass filter by substituting resistors for all the capacitors and capacitors for all the resistors. The circuits of Figs. 11 and 12 correspond to the low-pass filters in Figs. 9 and 10, respectively. Their characteristics are similar except that they are high-pass in nature. Actually, for high values of Q, the frequency responses of band-pass, low-pass and high-pass filters are pretty much the same close to the peak frequency. It's only when you get well away from the passband that you start to notice differences in attenuation.

Fig. 13 is a band-pass filter that uses the internal frequency compensation of the op-amp to replace one of the capacitors in the feedback network. This circuit has very high gain at low frequencies. Even at 50 kHz, the tuned n-l amplifier shown has a gain of about 200, which requires careful attention to circuit layout to

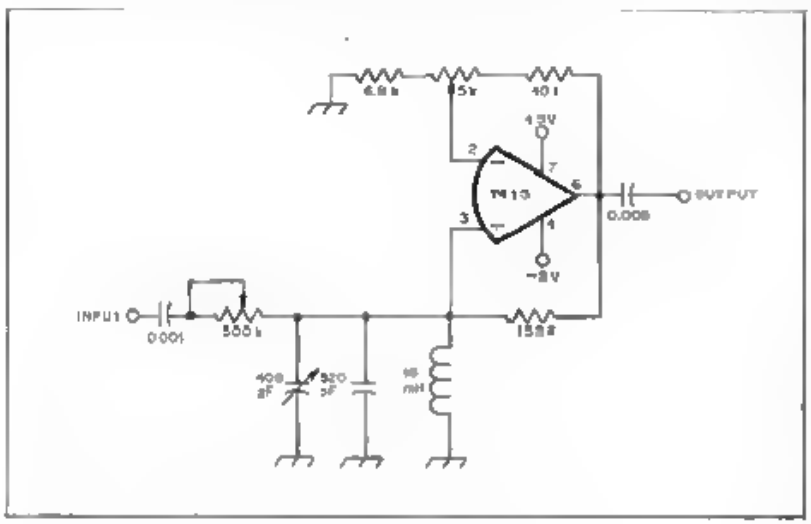

Fig. 13 — A tunable 50-kHz amplifier patterned after the circuit of Fig. 14.

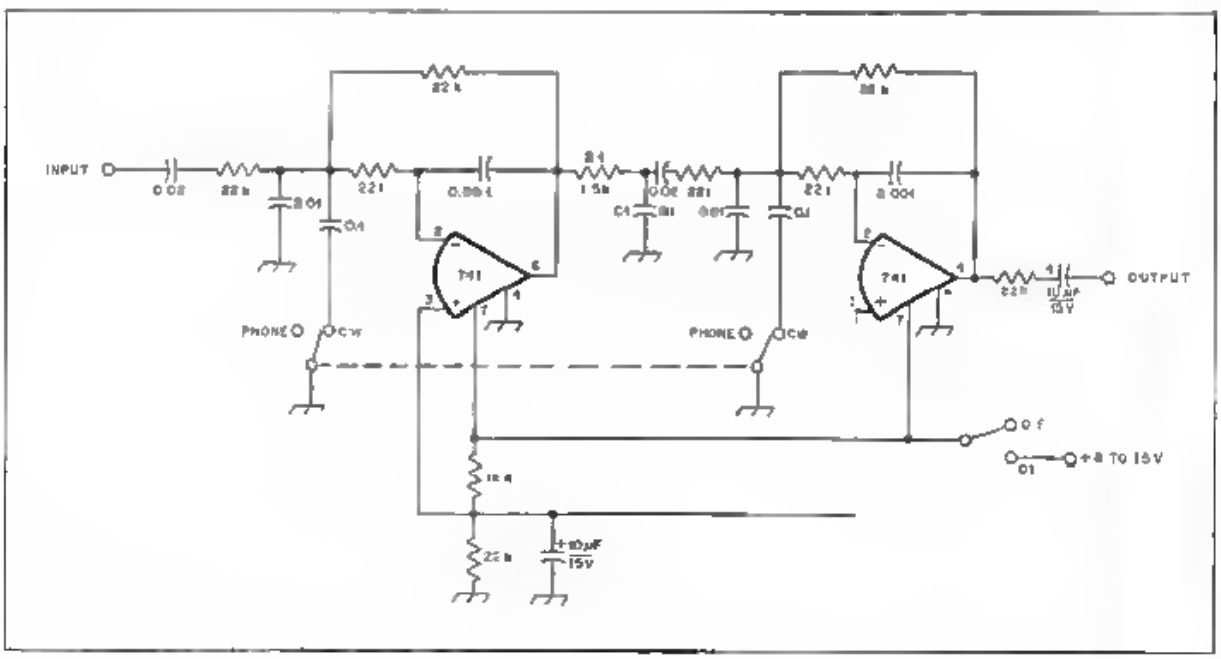

Fig. 16 — A combination phone and cw audio filter.

ensure stability.

LC Active Filters

One of the big advantages of active filters is that you can build high-Q filters without coils. On the other hand, if you *like* coils, you can still use them in active filter design. In fact, this will sometimes result in a more stable and reliable circuit. Fig. 14 is an example. This band-pass filter circuit increases the effective Q of the coil by means of positive feedback through R5. You can set the Q by adjusting R3. In this circuit, changing the bandwidth does not alter the gain. When properly adjusted, this filter is more stable and easier to use than some RC circuits, especially at high frequencies.

For example, you can build a practical 50-kHz tuned amplifier (Fig. 15) that is less critical to construct than one based on an RC design. My 10-mH coil had a measured Q of only 39 at 50 kHz, but it was easy to obtain bandwidths less than 370 Hz, indicating an effective Q of over 130. To align this filter, disconnect the input and adjust the 5-kΩ potentiometer until the circuit is on the verge of oscillation with the variable capacitor adjusted for the desired center frequency. With the input reconnected, the filter should be unconditionally stable.

Cascading Active Filters

Cascading passive filters can create problems, in that connecting the output of one filter to the input of another rather the impedances to interact, affecting the frequency response in ways you might not expect. Cascading active filters, however, is easy because the high-impedance input of each op amp doesn't affect the low impedance output of the preceding stage. The total frequency response is the product of the responses of the individual filters — that is, the total attenuation (in dB) at any frequency is the sum of the attenuations of the individual stages. Cascading filters greatly improves the stop-band attenuation. For example, if one filter has 20-dB attenuation at some frequency, two such filters in cascade will have 40 dB, three filters will have 60 dB, and so on.

Let's Build One

Enough theory; let's build one! Fig. 16 shows a useful circuit consisting of a pair of cascaded filters of the type described in Fig. 10. With the switch in the "phone" position, each section is a 2300-Hz low-pass filter with a Q of about one. R1 and C1 were added to further reduce the high-frequency response. Switching to cw adds extra capacitance, which not only lowers the resonant frequency to about 800 Hz, but also raises the Q to about 3.5. The two 0.02-μF coupling capacitors roll off the frequency response below 300 Hz, which helps to block any hum present on the input. The frequency responses for both modes are plotted in Fig. 17.

The filter may be driven by any audio source having less than about 2 kΩ output impedance and a voltage swing less than about 8 volts pk-pk on phone and 1 volt pk-pk on cw. (The gain is about one on phone and about eight on cw.) The output is sufficient to drive headphones of any impedance, but you should add an amplifier to drive a speaker.[1]

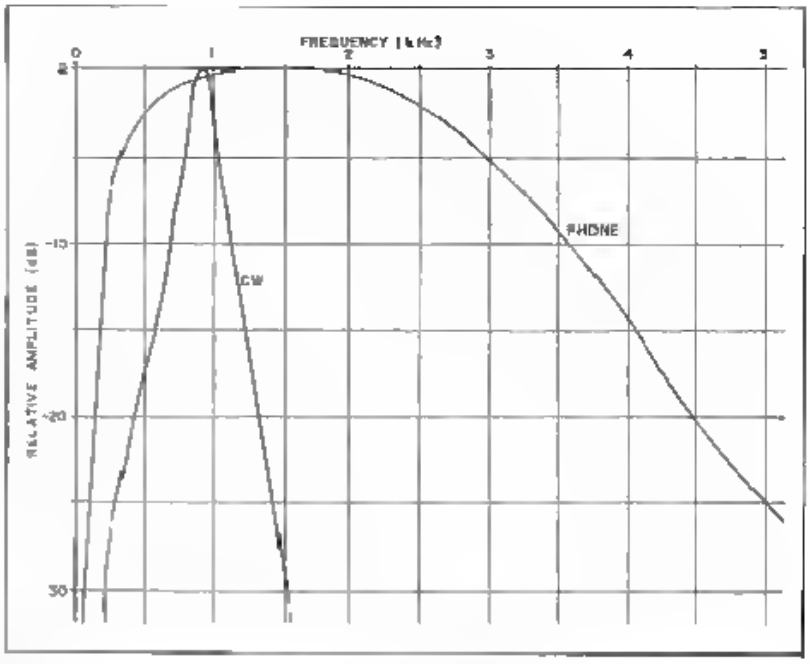

Fig. 17 — Frequency response of the phone/cw audio filter. In the phone mode the frequency response is 250 to 3100 Hz with a measured gain of 0.85. On cw, the gain is about eight with a 6-dB bandwidth of 300 Hz centered at 920 Hz.

By the way, it's not necessary to use two separate integrated circuits to build this filter. You can buy ICs with two or even four op amps in the package. For example, the Motorola MC1747 and MC4741 are the dual and quad versions of their MC1741 operational amplifier.

I hope this article has given you some idea of what can be done with active filters. In fact, there isn't much gain in the average ham shack where one of these little gizmos *wouldn't* come in handy. Drop one into your next construction project and see!

Notes

Woodward, "A Beginner's Look at Op-Amps," *QST*, April 1980, p. 15 and June 1980, p. 25.
Norris, "Adjustable Audio Filter for CW," *Ham Radio*, August 1980.
Mumme, "A Handy Audio Amplifier," Hints and Kinks, *QST*, December 1979, p. 16.

References

Bridle, *Passive and Active Network Analysis and Synthesis*, Houghton Mifflin, 1974.
Lam, *Analog and Digital Filters*, Prentice-Hall, 1979.

A Simple, High-Performance CW Filter

By Ed Wetherhold, W3NQN
1426 Catlyn Pl
Annapolis, MD 21407

This inductor-capacitor CW filter uses one stack of the familiar 88-mH inductors and two 44-mH inductors in a five-resonator circuit that gives high performance at low cost. The center frequency is fixed at 750 Hz because most transceivers use this sidetone frequency, but sidetones between 700 and 800 Hz can be received with less than 1 dB attenuation relative to the center frequency. Ed Wetherhold, W3NQN, designed and built the filter presented here. The author can provide parts for this project at nominal cost. Write E. E. Wetherhold, W3NQN, 1426 Catlyn Place, Annapolis, MD 21401 for more information. If you need a design for a different center frequency, the author can provide that as well. Be sure to include a self-addressed, stamped 9½- × 4-inch envelope with your request.

One feature of this filter is a 3-dB bandwidth of 236 Hz. This bandwidth is narrow enough to give good selectivity, and yet broad enough for easy tuning with no ringing. Five high-Q resonator circuits provide good skirt selectivity that is equal to or better than most commercial active filters costing more than $60. In comparison, this CW filter can be built for less than $15. Simple construction, low cost and good performance make this filter an ideal first project for anyone interested in putting together a useful station accessory.

Design

Fig 1 shows the filter schematic diagram and component values. These values were selected for a center frequency of 750 Hz and for a filter impedance level of 230 ohms. The filter sees a 230-ohm source impedance consisting of the 200-ohm source (transformed from 8 ohms), a 22-ohm transformer winding resistance and an 8-ohm inductor resistance. In a similar way the filter sees a load impedance of 230 ohms. This design was selected so that only one turn needs to be removed from both windings of a standard 44-mH inductor to give the required L2 and L4 values.

Construction

Fig 2 is a pictorial diagram showing the filter wiring. Note the 44-mH lead connection, as well as the connections between the capacitor leads, the 88-mH stack terminals and the 44-mH inductor leads. Fig 3 shows the finished filter installed in an aluminum box. Before beginning construction, obtain one 88-mH five-inductor stack with a mounting clip and two 44-mH inductors, and then follow steps 1 to 5.

1) Remove one turn from each of the two windings of one 44-mH inductor to get 43.5 mH (total turns removed is two). Carefully scrape off the film insulation and connect the start lead (with sleeve) of one

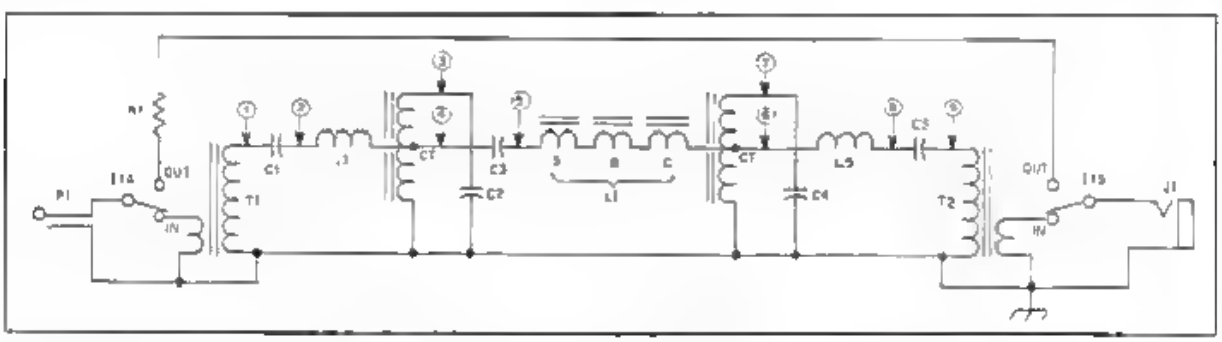

Fig 1—Schematic diagram of 750-Hz CW filter. Use 1% tolerance capacitors for best results.

C1–C5—0.512 µF capacitor
C2, C4—1.036 µF capacitor
C3—170.7 nF capacitor
J1—Phone jack or jack to match your headphones
L1–L5—88 mH toroid (part of toroid stack, see text)
L2, L4—43.5 mH toroid (modified 44-mH toroid, see text).
L3—264-mH toroid (part of toroid stack, see text).
P1—Phone plug or plug to match your receiver
R1—Zero to 220-ohm, ½-W 10% resistor (see text)
S1—DPDT switch
T1, T2—8-ohm to 200-ohm impedance-matching transformer, 0.4 W

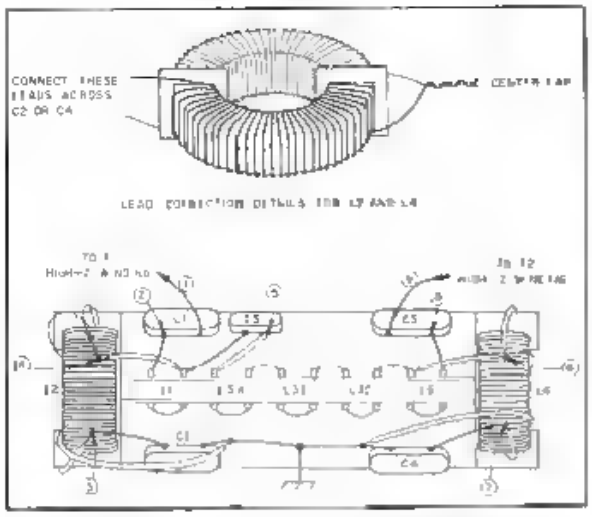

Fig 2—Pictorial diagram showing L2 and L4 lead connection and wiring of inductor stack (L1, L3, L5).

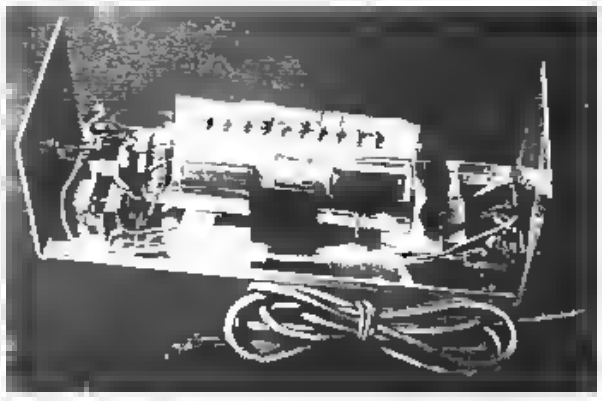

Fig 3—The assembled filter is shown installed in a CR-800 aluminum box. The filter/bypass switch (S1) and input/output transformers (T1, T2) are at the left end of the box.

winding to the finish lead (no sleeve) of the adjacent winding to make the center tap as shown in Fig 2. Do the same for the second 44-mH inductor.

2) Fasten both of the 43.5-mH inductors to opposite ends of the 88-mH stack using clear silicone-rubber sealant, available from most hardware stores.

3) Position the 43.5-mH inductors so their leads can be easily connected to the rest of the circuit. Solder the capacitor leads to the stack terminals as shown in Fig 2.

4) Obtain a suitable box and make holes for the inductor mounting clip, the DPDT switch, and the phone jack and phone cord. First, install matching transformers T1 and T2 and the inductor stack with capacitors. Fasten the transformers (with leads pointing up) to the bottom of the box with silicone rubber sealant. Secure the stack to the bottom of the box with a 1-3/8-inch component mounting clip and two no. 6-32 × 5/16-inch screws. Instead of the 8 × 3 × 2¼-inch aluminum box shown in Fig 3 (Mouser Stock No. 537-CR-800), a small cardboard box may be used to minimize cost.

5) Complete the wiring of the transformers, the DPDT switch with resistor R1, and the phone jack and phone plug. Then check the correctness of your wiring by measuring and comparing the filter node-to-node resistances with the values listed in Table 1.

Table 1

Node-to-Node Resistances for the CW Audio Filter

Nodes		Components	Resistance
From	To		(ohms)
1	GND	T1 hi-Z winding	12
2	GND	L1 and ½ L2	10
3	GND	L2	4
4	GND	½ L2	2
5	GND	L3 and ½ L4	26
6	GND	½ L4	2
7	GND	L4	4
8	GND	L5 and ½ L4	10
9	GND	T2 hi Z winding	12
2	4	L1	8
5	8	L3	24
6	8	L5	8
2	3	L1 and ½ L2	10
6	7	L5 and ½ L4	10

Notes

1) See Figs 1 and 2 for the filter node locations.

2) Check your wiring using the resistance values in Table 1. If there is a significant difference between your measured values and the table values, you have a wiring error that must be corrected.

3) For accurate measurements, use a digital VOM or an analog VOM (such as a Triplett Model 630) that has a scale center of about 5 ohms on its ×1 ohmmeter range.

Installation

T1 and T2 match the filter to the receiver low impedance audio output and to an 8 ohm headset or speaker. If your headset is high impedance, T2 may be omitted. In this case, connect a 10%, ½ W resistor from node 9 (C5 output lead) to ground. Choose the resistor value so the parallel combination of the headset and resistor gives the correct filter termination impedance (within 10 percent of 250 ohms).

Performance

The measured 30-dB and 3-dB bandwidths are about 511 and 235 Hz, respectively, and the 30-dB/3-dB shape factor is 2.17. This factor can be used to compare the performance of this filter with others. The measured insertion loss at 750 Hz is less than 3 dB and is typical of passive filters of this type. This small loss is compensated by slightly increasing the receiver audio gain. R1 helps to maintain a constant audio level when the filter is switched out of the circuit. The correct value of R1 for your audio system should be determined by experiment. Start with a short circuit for R1 and then gradually increase the resistance until the audio level appears to be the same with the filter in or out of the circuit.

More than 700 hams have constructed this five-resonator filter (using either the 2-stack or the newer 1-stack arrangement) and many have commented on its excellent performance and lack of hiss and ringing.

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Wetherhold, "Modern Design of a CW Filter using 88- and 44-mH Surplus Inductors," *QST*, Dec. 980 and Feedback, *QST*, Jan 1981, p. 43.

Wetherhold, "High Performance CW Filter," *Ham Radio*, Apr 1981.

Radio Handbook, 23rd edition, W Orr, ed., Howard W. Sams & Co, 1987 (1-Stack CW Filter), p. 13-4.

From 1990 ARRL Handbook *Ch 28, p 3:*

A Passive Audio Filter for SSB

By Ed Wetherhold, W3NQN
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Annapolis, MD 21407

While audio filters are most often used during CW reception, the SSB operator can also benefit from their use. Shown in Figs 4 and 5 is a passive band-pass filter designed by Ed Wetherhold, W3NQN, for phone operation. This filter was described in Dec. 1979 *QST*.

All of the inductors are the surplus 88-mH toroidal type with their windings wired either in series or parallel to get the required 88 or 22 mH of inductance. The series connection is shown in Fig 2. The 0.319-μF capacitors were selected from several 0.33-μF capacitors that were about 3 percent on the low side. The 0.638-μF value was obtained with a single 0.68-μF capacitor that was about 6 percent on the low side. The 1.276-μF values were obtained by paralleling selected 1-μF and 0.33-μF capacitors.

Fig 5 shows the measured and calculated attenuation responses of the filter. The difference between the measured and calculated response at the low frequency side of the passband is probably caused by the much lower Q of the inductors at these frequencies.

The necessary termination resistance of this filter is 206 ohms. While this is not a standard value, it should not be too difficult for most amateurs to accommodate. If low impedance headphones are used, a matching transformer can be used to provide the correct termination. A suitable transformer is available from Mouser Electronics (see Chapter 35 parts-suppliers list). The part number is 42TU200, and it is a 200-ohm CT to 8-ohm CT unit.

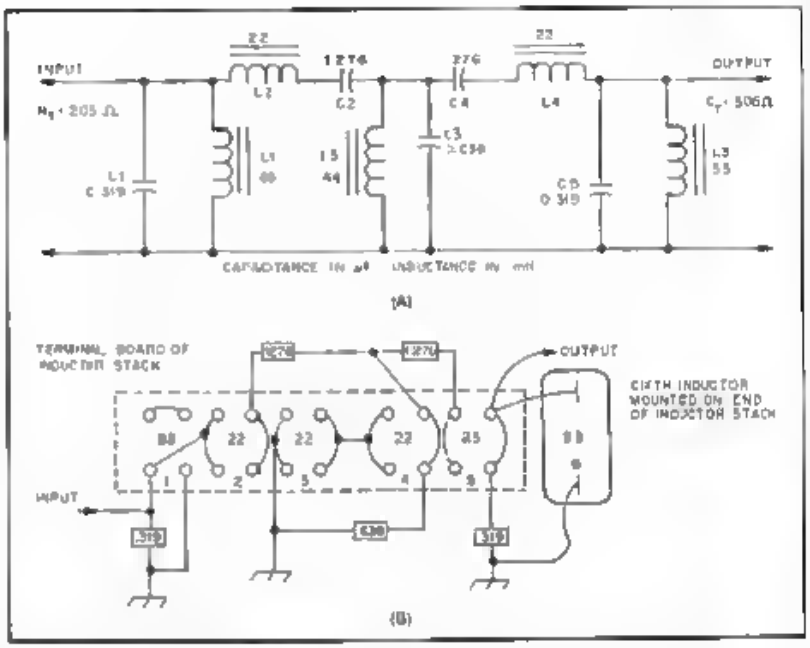


Fig 4—Schematic diagram of the SSB band-pass filter (A). Shown in B is a pictorial wiring diagram of the terminal board on the inductor stack.

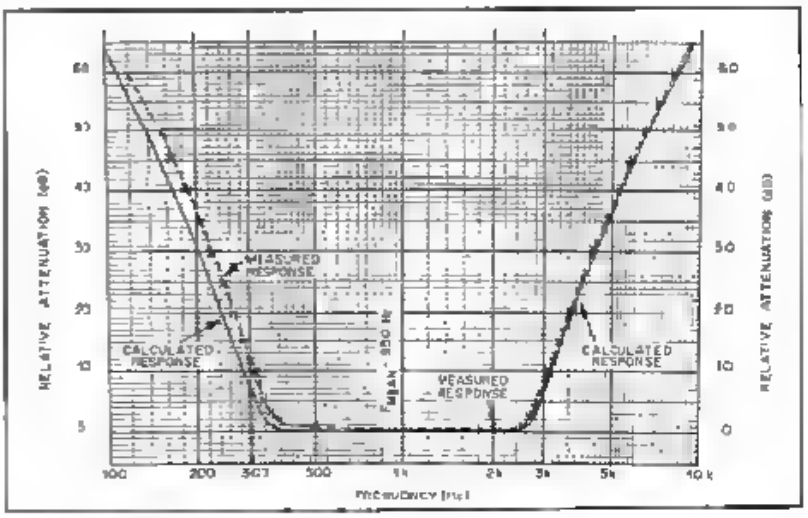


Fig 5—Response curves of the SSB band-pass filter.

From July 1987 QST, p 24

Designing and Building Simple Crystal Filters

A simple and inexpensive crystal filter that performs well makes receiver and transmitter projects much more fun. Build one yourself at a fraction of the cost of a commercial unit.

By Wes Hayward, W7ZOI
7700 SW Danielle Ave
Beaverton OR 97005

I am encouraged by the large number of radio amateurs who want to build their own rigs. The ready availability of good-quality semiconductors helps to this pursuit. Other components are sometimes harder to find, at least at an affordable price. One example is the crystal filter—the heart of any superheterodyne receiver or transmitter.

Inexpensive crystals are readily available. They should be characterized and matched for frequency prior to use in a typical crystal filter. Methods for building the needed test equipment and performing the measurements have been presented before.[1] These methods are, unfortunately, somewhat complicated for the casual experimenter who may hesitate to construct special test equipment when just one filter is to be built. What experimenters really need is an empirical filter design method, one that lends itself to casual "tweaking." Such a method is described in this article.

The Cohn Filter

In the course of computer studies of both crystal and LC filters, I've noted that a circuit called the "Cohn," or "Mn-loss" filter, lends itself to particularly simple designs.[2] This filter configuration derives its name from its originator, and differs from the more familiar Butterworth and Chebyshev circuits. The Butterworth bandpass filter is built for optimum flatness at the filter center. The Chebyshev design allows equal passband ripples, and is designed for the best stopband attenuation (steepest skirt response). The Cohn filter is a compromise. It is optimized to exhibit minimum insertion loss when built with practical resonators, while preserving a good shape factor. The Cohn filter, in LC form, is not new to the radio amateur.[3,4] It is not limited to LC resonators, however. It works great with crystals!

The Cohn filter, crystal or otherwise, is a rather simple circuit. This becomes more apparent when we view the filter using coupled-resonator methods.[5] All normalized coupling coefficients are equal. Moreover, the normalized end-section loaded-Q factor is the reciprocal of the coupling coefficient. The practical simplification becomes apparent if we examine the generalized crystal filter circuit shown in Fig 1. All capacitors in the circuit are of equal value! The shunt capacitors are coupling elements while the series capacitors in the filter end sections are included to properly tune the circuit.

Practical Cohn Crystal Filters

An empirical method that the amateur may use for crystal filter design is described easily in a step-by-step procedure.

1) Obtain a collection of substantially identical crystals. The crystals are first matched in frequency. The same oscillator should be used to measure all crystal frequencies. The error (frequency difference) should be less than 10% of the *desired bandwidth* of the filter. For example, a filter with a 1-kHz bandwidth should use crystals matched to within 100 Hz or better.

2) Pick a capacitance value to be used in the filter. The capacitance (C) value determines the filter bandwidth. Larger C values yield narrower bandwidth and higher insertion loss.

3) Vary the end terminations to obtain a shape that is free of passband ripple while

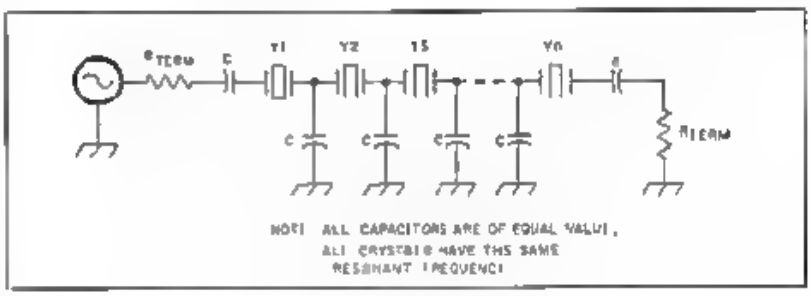


Fig 1—Generalized crystal filter suitable for empirical construction.

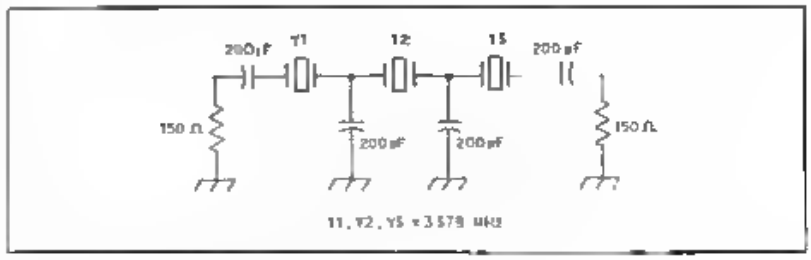


Fig 2—A simple CW filter using three crystals.

*Notes appear at end of article

providing sufficient stopband attenuation.

This empirical procedure is illustrated in the following examples. I've cheated a bit—I used a personal computer to simulate the filter, and generate the data presented, but I've obtained similar results with filters I have built. The experimental results agree well with the computer models. All examples shown are based on a collection of crystals from my junk box. They are inexpensive 3.579-MHz TV color-burst crystals. The average motional inductance for these crystals is 117 mH, with a (rather poor) typical Q of 50,000. The parallel capacitance is about 4 pF.

A Three-Crystal Cohn Filter

A simple and practical filter for a beginner's first CW superheterodyne receiver is shown in Fig 2. Three crystals are used. The capacitors are 200-pF units, a standard value. Experimentation (done here with the computer) shows that a good filter shape is obtained with an end termination of 150 ohms. Fig 3 shows the frequency response of this filter. The −3 dB bandwidth is 403 Hz, and the insertion loss is 3.8 dB. The loss will be lower with better (higher Q) crystals. The impedance match is shown in the figure as a series of dots. This is the return loss normalized to the source impedance—150 ohms for the filter shown.

If different crystals are used, the same bandwidth can still be obtained, within limits. The coupling capacitors and end terminations will then be different, however. Insertion loss will also differ.

Decreasing the value of the capacitors increases the bandwidth. Some practical values are shown in Table 1, again the result of tweaking with the computer. This will provide some guidance in experimentation.

Fig 4 illustrates the effect of altering the terminating resistance. Fig 4A shows the result of 75-ohm terminations, lower than the desired 150-ohm value. The filter shows some passband ripple and a higher insertion loss. The effect of a 300-ohm termination is shown in Fig 4B, where the peak shape becomes more rounded, with degradation of skirt response. While the poorer frequency domain shape is generally less desirable, the filter with the higher termination has a significantly improved group delay; this filter would be preferred for high-speed data applications.

A Six-Crystal Cohn Filter

The three-pole filter mentioned above is practical. It does not, however, offer skirts that are as steep as we would like for many demanding applications. Improved skirt selectivity in a filter is obtained by using more crystals. The computer can be used to generate another table like that shown for the three-crystal filter. Alternatively, the results of Table 1 can be used as a starting point for experimentation. The

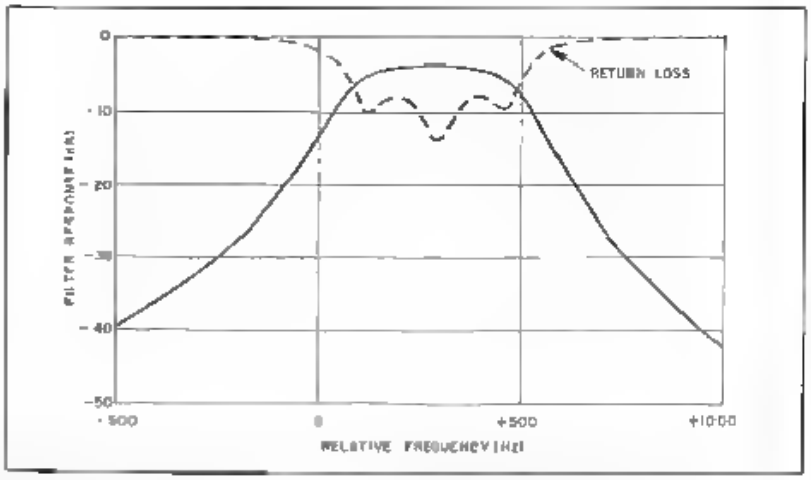


Fig 3—Frequency response of the filter of Fig 2. The dots show the input return loss, indicating the quality of the impedance match.

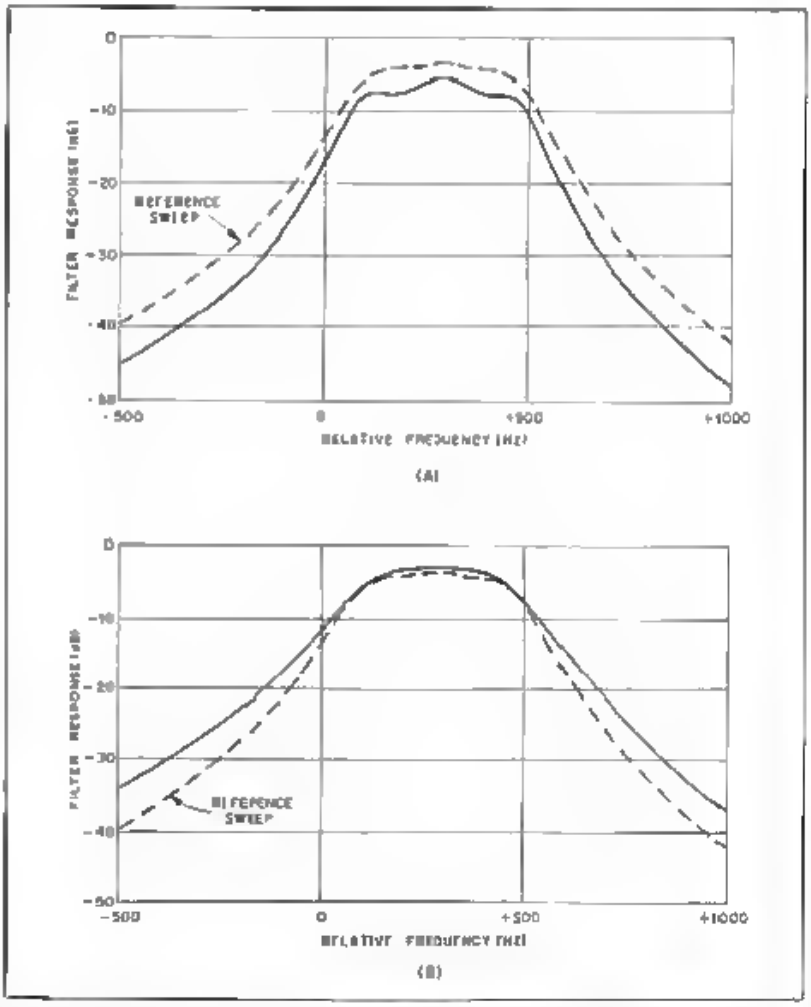


Fig 4—Frequency response of the filter of Fig 2 with changes in the end terminations. Curve A shows response with 150-ohm terminations, curve B shows the response using 300-ohm resistors. See the text for considerations of which is "better."

ARRL Lab Experiments with the Cohn Filter

ARRL Lab staff members were intrigued by the material on Cohn filters presented by Wes Hayward, W7ZOI. We built four CW filters and one SSB filter, following Wes's instructions. Tests confirmed the computer models developed by Wes. This was no surprise!

CW Filters

Four different batches of crystals were used for the CW filters. The crystal sources were identifiable, and the relative quality of each batch was determined. Four filters were constructed (Fig A). With the exception of the crystals used in each filter, the filters were identical. The filter schematic is shown in Fig B. The capacitors are 300-pF, 5%-tolerance silver-mica types. The 500-ohm terminations (variable resistors) at the ends of the filter were used to "trim" the filter for the best shape and response characteristics during testing. An HP-8540 spectrum analyzer was used to generate the filter response curves shown in photos C through G.

The units used in filter no. 1 are TV color-burst crystals (3.579545 MHz). They were purchased originally from Radio Shack (about $1.60 each) for another project. There were only five of these crystals in the batch, so frequency matching (within 50 Hz) was not as close as with some of the other crystal batches.

The crystals used in filter no. 2 were selected from an assortment of ten 4.000-MHz microprocessor units purchased from JAN Crystals.[1] These crystals were frequency matched within 40 Hz. The crystals cost approximately $3 each.

Filter no. 3 uses crystals selected on the basis of frequency matching from a large batch (over 30) of 4.000-MHz microprocessor crystals on hand in the ARRL Lab (matched within 30 Hz). These crystals can be characterized as "grab bag" quality, and similar units are available from various dealers at a cost of less than $1 each.

We bought the crystals used in filter no. 4 from International Crystal Co.[2] They can be characterized as high-quality, moderate-cost units. Their guaranteed frequency tolerance is 0.001% of 4.000000 MHz, matching was within 8 Hz, and cost is approximately $10 each.

SSB Filter

A four-crystal 12-MHz SSB filter was built using 160-pF 10%-tolerance silver-mica capacitors. An 8:5 transformer is used for impedance matching. The crystals are microprocessor types purchased from Jameco Electronics[3] at a cost of approximately $1 each. Of 12 crystals purchased, only 10 were suitable for filter use. The filter response is shown in photo G.

Test Results

Photos C through F show the response curves of the four CW filters. Photo G shows the response curve for the 12-MHz SSB filter. Insertion loss is quantified only for CW filter no. 4 because series resistors were used to adjust the terminating impedance of filter nos. 1 through 3. These resistors introduce losses. In practice, each filter would be coupled to its associated circuitry through matching transformers, not resistors.

Filter no. 1 exhibits an extremely sharp response with a bandwidth of approximately 240 Hz at the −3 dB points; it may be too sharp for good *CW* copy. Changing the 300-pF capacitors in this filter to a lower value will broaden the response.

Filter no. 2 is not quite as sharp as filter no. 1, and exhibits a peak ripple effect. The response asymmetry can be corrected by trimming the filter.

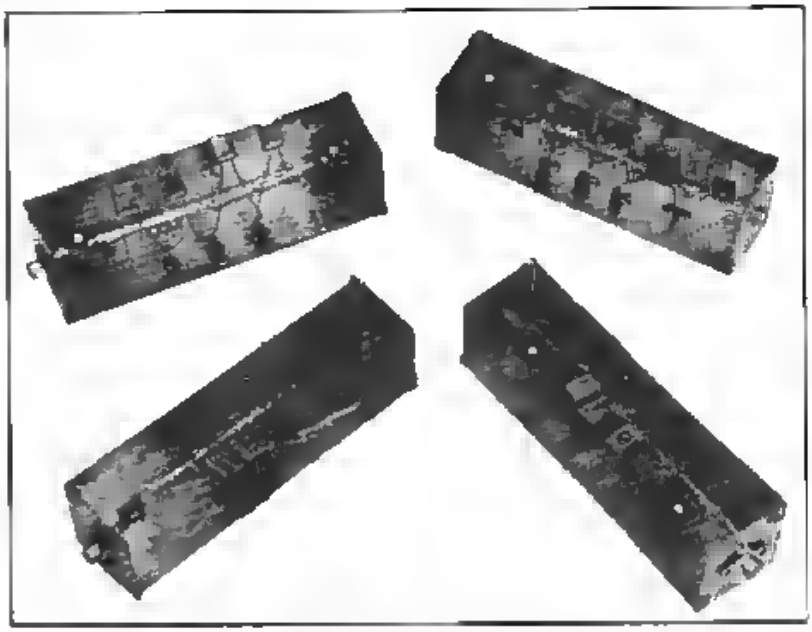


Fig A—Four CW crystal filters were built in this configuration. The PC-board mounting surfaces provide a ground plane. Capacitors are soldered directly to the ground plane, and the crystals are connected using the capacitors as standoffs. Photo jacks are used for input and output connectors. The only variables in the construction of the filters are the crystal characteristics and the length of the crystal leads. The SSB filter is not shown.

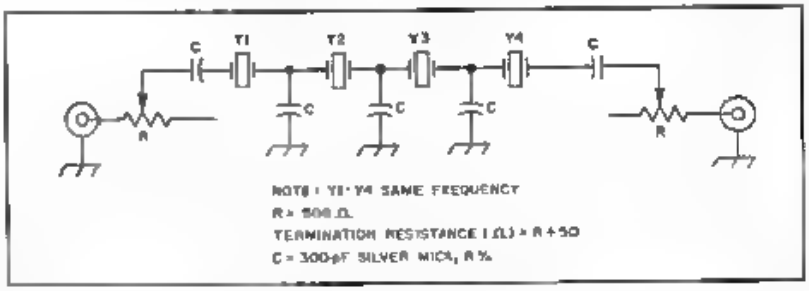


Fig B—Schematic diagram of the crystal filters. Capacitors are all of equal value. Terminating resistors are variable 500-ohm units. Crystals are all of equal nominal frequency with minor (up to 50-Hz) variation.

[1]JAN Crystals, 2400 Crystal Dr., PO Box 06017, Fort Myers, FL 33906-6017, tel. 813 936-2397.

[2]International Crystal Manufacturing Co., Inc., PO Box 26330, 701 W. Sheridan, Oklahoma City, OK 73126-0330, tel. 405 236-3741.

[3]Jameco Electronics, 1355 Shoreway Rd., Belmont, CA 94002, tel. 415-592-8097.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

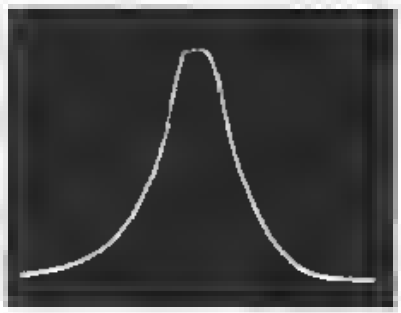


Fig C—Spectral photo showing the response of filter no. 1. Horizontal divisions are each 200 Hz; vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 3.579 MHz.

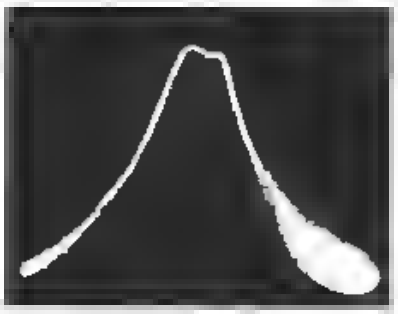


Fig D—Spectral photo showing the response of filter no. 2. Horizontal divisions are each 200 Hz, vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

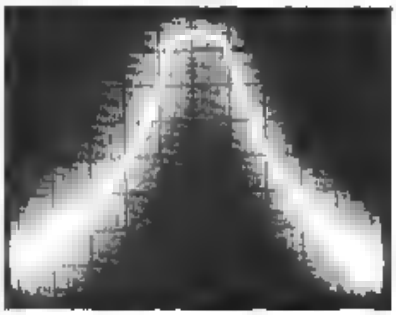


Fig E—Spectral photo showing the response of filter no. 3. Horizontal divisions are each 200 Hz, vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

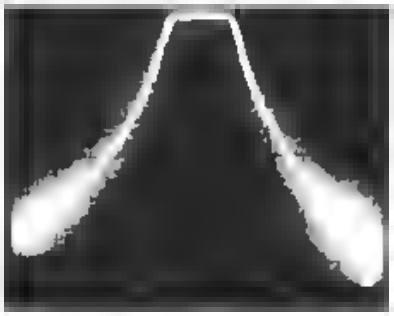


Fig F—Spectral photo showing the response of filter no. 4. Horizontal divisions are each 200 Hz, vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 4.000 MHz.

with the termination resistors, but insertion loss increases significantly. For CW use, however, this filter is probably more than adequate.

Filter no. 3 shows a bandwidth of approximately 360 Hz at the −3 dB points. It is symmetric and shows low ripple. This is a very good CW filter.

Filter no. 4 is a good example of what can be accomplished with high-quality crystals and proper terminations. This filter is used in a CW receiver designed by Dave Newkirk, AK7M (see cover of this issue). The input and output impedances of this filter are 200 ohms. To match the 50-ohm impedance of the test setup 4:1 transformers were used. Filter insertion loss is 2 dB, with an ultimate rejection of over 90 dB.

The SSB filter shows a −3 dB bandwidth of approximately 2.1 kHz. There is no discernible ripple, and the insertion loss is 4.4 dB.

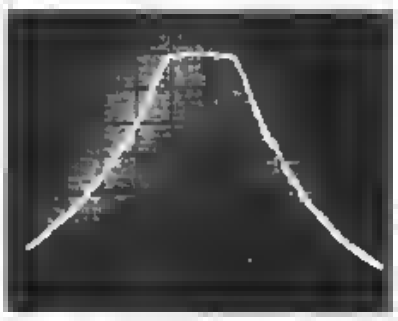


Fig G—Spectral photo showing the response of the SSB filter. Horizontal divisions are each 1 kHz, vertical divisions are each 10 dB. Sampling bandwidth is 100 Hz. The center frequency is 12.000 MHz.

Conclusions

The empiric approach to designing Cohn filters for CW or SSB use is a viable alternative to purchasing commercial filters. The relatively high component cost for the best filter design tested (CW filter no. 4) still results in an advantage of over 50% when compared to the price of commercial equivalents. All of the filters tested are adequate for most home-brew projects. They are fun to build, and result in appreciable savings.—*Bruce O. Williams, WA6IVC, ARRL Staff*

Table 1

Cohn Three-Crystal Filter

Bandwidth (Hz @ −3 dB)	*C* (pF) (k = 1000)	R_{end} (Ohms)
360	200	150
600	130	238
1.1k	70	431
1.6k	30	1.5k
2.5k	17	3.3k

computer was used in the "construction" of a filter with six crystals. The circuit, again a narrow CW filter, is shown in Fig 5. The 200-pF capacitors used in the earlier filter are retained. The frequency response of this six-crystal filter is shown in Fig 6, where the "reference sweep" is the response of the previous three-element filter. The new filter has a −3 dB bandwidth of 354 Hz, but much steeper skirts than the three-element filter.

A Simple SSB Filter

Table 1 shows a number of simple three-pole filter configurations. Bandwidth is increased for a given set of crystals merely by decreasing the capacitance value. The frequency domain response for a three-pole SSB filter with 30-pF capacitors is shown in Fig 7. The "reference sweep" is the response of the earlier three-pole CW filter with 200-pF capacitors. The skirt response of the SSB three-crystal filter is certainly less than spectacular. More crystals will improve this response significantly. This simple three-pole filter is still practical for some applications, however, such as a portable VHF SSB transceiver.

Experimental Methods

The computer-based "experiments" have proved to be useful. There are generally no surprises. I've "built" filters on the computer using more than a dozen crystals. Some of the more practical designs have been transferred to hardware for receiver applications. Many of these designs operate at different frequencies, some using 4.433-MHz European TV color-burst crystals. These crystals are harder to obtain, but their frequency is more compatible with the existing HF ham bands, avoiding the spurious responses that can sometimes occur with a 3.579-MHz IF.

Almost all of my test equipment is built for an input and/or output impedance of 50 ohms. The test equipment is still easily used for filter experiments. Extra resistance is merely added at the filter input and output to bring the level up to that desired. This is illustrated in Fig 8. Ferrite transformers may also be built to transform impedance levels, but they cannot be changed as quickly as resistors.

It is often convenient to experiment with a filter that is contained within a receiver or transmitter. An example is shown in the

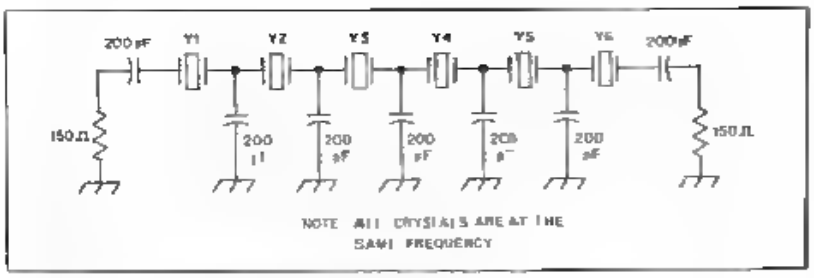


Fig 5—Circuit of a Cohn filter using six crystals

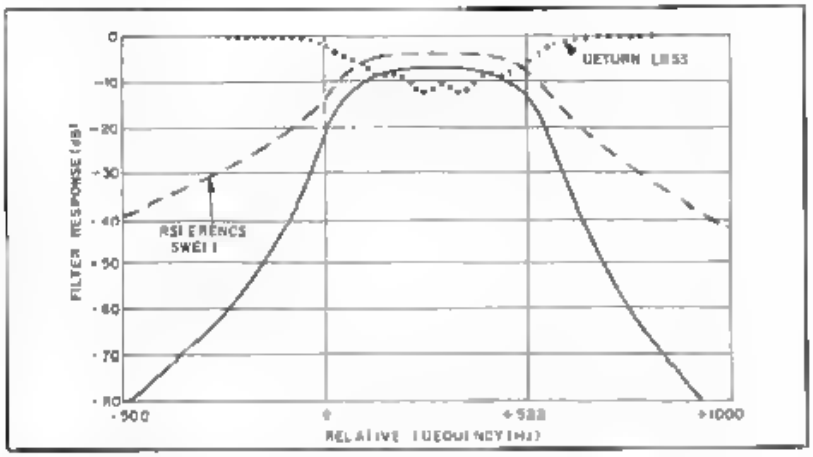


Fig 6—Frequency response of the six-crystal filter. The reference sweep is the response of the three-crystal filter of Fig 2.

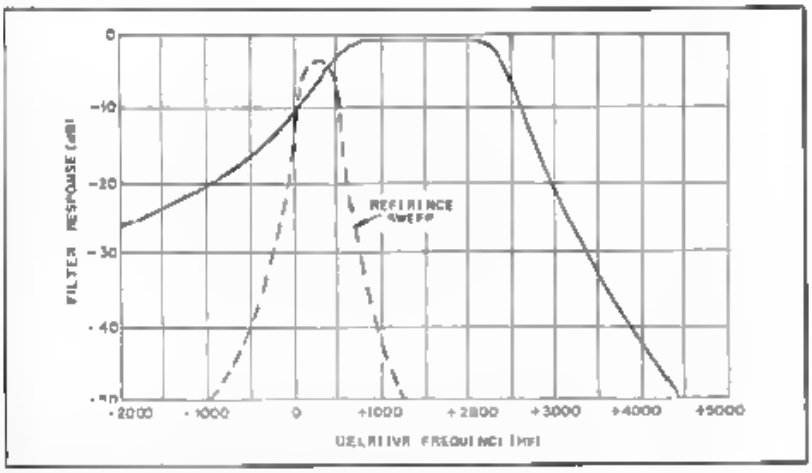
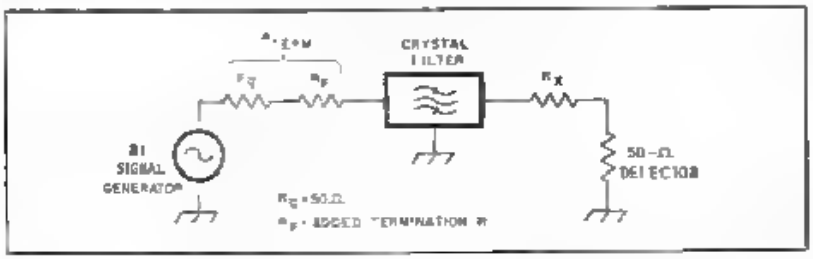


Fig 7—Frequency response of a simple three-pole SSB filter. The circuit is that of Fig 2 with all capacitors changed to 30 pF and terminations of 1500 ohms.



← Fig 8—External resistors may be added to an experimental filter to allow use of 50-ohm instrumentation for circuit evaluation.

partial schematic of Fig 9A. Q1 is a dual gate MOSFET mixer. The drain resistor determines the input loading impedance for the filter. An identical resistor terminates the filter output. An NPN amplifier, Q2, buffers the output—insurance that the following stages will not alter the crystal filter termination. Fig 9B is a modified form of the same filter. Tuned circuits have been inserted to present higher impedance to the transistors, affording more gain. The output amplifier is changed to a JFET. This modified circuit is better suited to higher impedance filters, as might be encountered with an SSB transmitter or receiver. Once the circuit containing the filter is built, filter response may be measured by tuning the receiver through a steady carrier while observing the output of a later stage with an oscilloscope or RF voltmeter.

It's often difficult to build a filter while also building a receiver. If problems occur, it's hard to tell if they are related to the filter or to the rest of the circuitry. Uncertainty is removed if receiver construction begins with a simpler, single-crystal filter. This allows you to get the receiver working before pursuing the better filter. I don't encourage you to retain the single-crystal filter as a final option. The enhanced performance afforded by additional crystals is more than ample justification for the minimal added effort and expense.

Other Crystals

The examples presented have used readily available color-burst crystals. There is nothing special about them. Indeed, they often represent the poorest possible quality for a crystal, and their frequency (3.579 MHz) can cause compatibility problems in many of the ham bands. They are, however, both available and cheap.

Many parts distributors list crystals for microprocessor applications in their catalogs. The only experience I have had with these crystals was with two 4 MHz crystals.[6] The average Q was 150,000, motional inductance was 148 mH and the two crystals differed in frequency by 165 Hz. Further data on other crystal types would be of great use to the amateur community. Anyone out there with data to share? [See the sidebar to this article. Ed.]

Traditional intuition might suggest that narrow-bandwidth filters are more difficult to design and build than those with wider bandwidth. Just the opposite is true; CW filters are easier to build than SSB or AM filters. This is fortunate, for it seems that much of the present home-brew activity is aimed at CW rigs.

Narrow-bandwidth CW filters are easily built with the lower frequency crystals, such

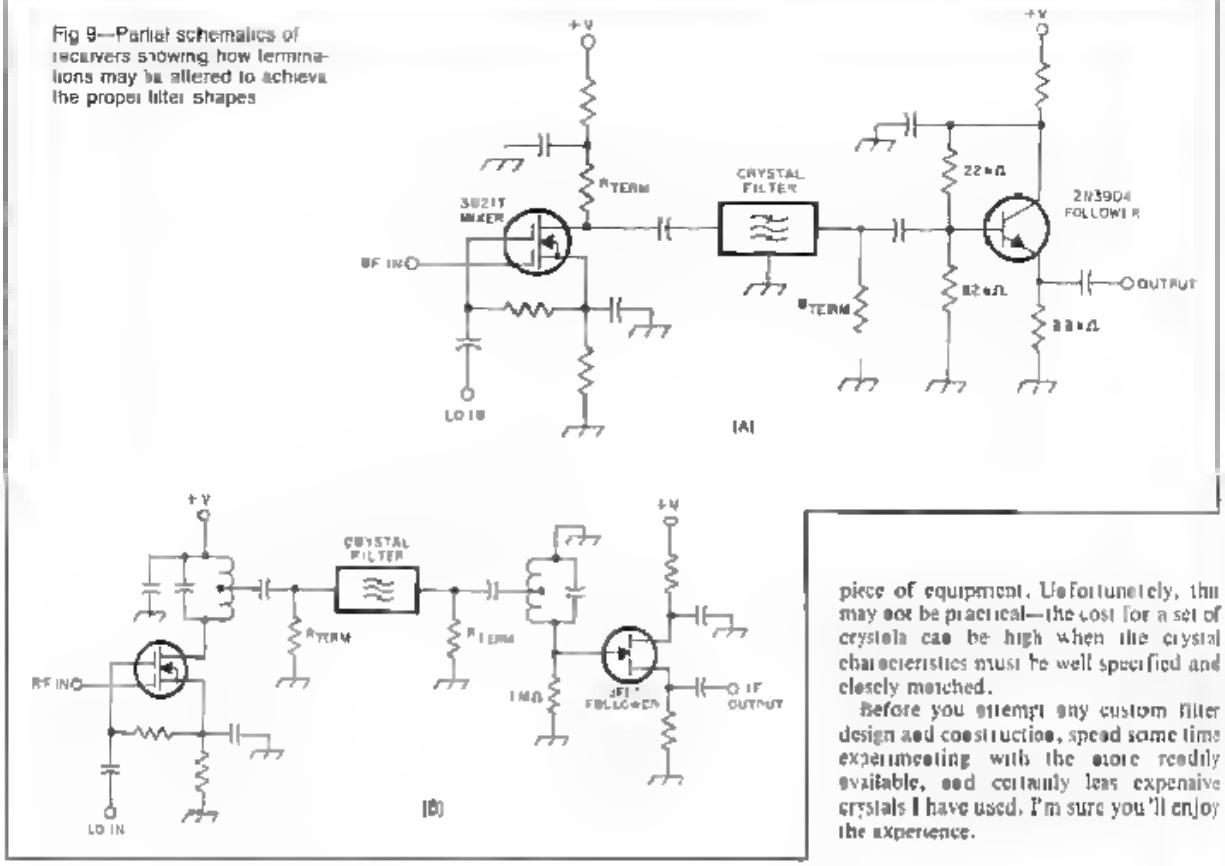


Fig 9—Partial schematics of receivers showing how terminations may be altered to achieve the proper filter shapes

as those at 3.579 MHz. While an SSB filter can be built at 3.579 MHz, probably higher terminating impedances will be required. The termination value drops with increasing frequency, making wider bandwidth filters more easily realized at higher frequencies. I often build equipment with a 10-MHz IF because crystals with excellent Q are readily available for this frequency.

Typical parameters for these crystals are: motional inductance = 20 mH, parallel C = 3 pF and Q = 200,000. These characteristics result in practical CW filters with terminating impedances as low as 50 ohms, and SSB filters with 200- to 500-ohm loads.

You can, of course, order high-quality crystals for any desired frequency. It is also possible to fit a new filter into an existing piece of equipment. Unfortunately, this may not be practical—the cost for a set of crystals can be high when the crystal characteristics must be well specified and closely matched.

Before you attempt any custom filter design and construction, spend some time experimenting with the more readily available, and certainly less expensive crystals I have used. I'm sure you'll enjoy the experience.

Notes

[1] W. Hayward, "A Unified Approach to the Design of Crystal Ladder Filters," *QST* May 1982, pp 21-27.

[2] S. Cohn, "Dissipation Loss in Multiple Coupled Resonators," *Proceedings IRE*, Aug 1959.

[3] W. Sabin, "The Solid-State Receiver," *QST* Jul 1970, p 35.

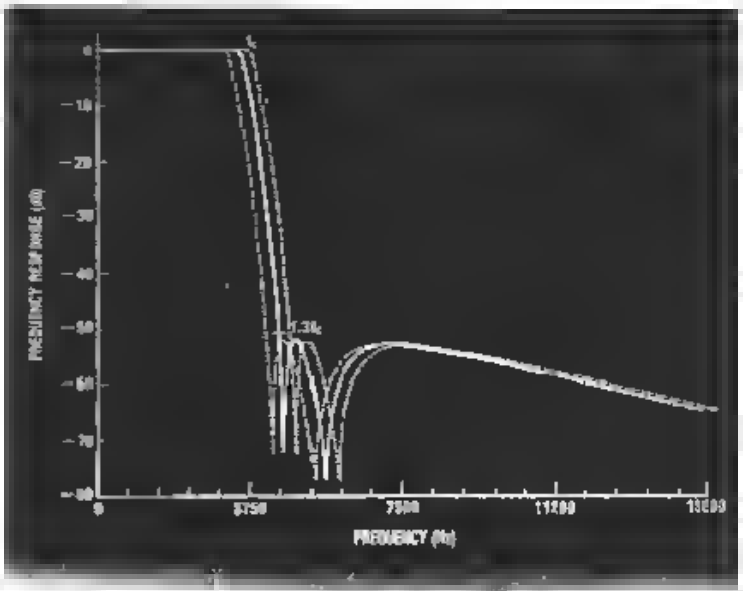
[4] D. DeMaw, "His Eminance—The Receiver," *QST*, Jun and Jul 1976.

[5] A. I. Zverev, *Handbook of Filter Synthesis* (New York: John Wiley and Sons, 1967).

[6] Meuser Electronics, 11511 Woodside Ave, Lakeside, CA 92040, part no. ME332-1040.

From April 1986 QST, p 13:

SuperSCAF and Son—A Pair of Switched-Capacitor Audio Filters



photograph courtesy Gould AMI Semiconductors, copyright 1986

Been looking for an audio filter that's a great performer and is easy to build? Here are two that fill the bill nicely!

By Rich Arndt, WB4TLM and Joe Fikes, KB4KVE

179 Wildwood Dr
Sanford, FL 32771

6917 Criner Rd
Huntsville, AL 35802

Nothing is more frustrating than trying to copy a weak signal in heavy QRM except, perhaps, losing it altogether. A good audio filter can be tremendously helpful in separating the weak signals from the strong ones. The two switched-capacitor filters (SCFs) presented here reflect the needs of different users. SuperSCAF is a self-contained audio filter with thumbwheel frequency selection at 100-Hz intervals, a built-in audio power amplifier and an ac-operated power supply. Junior SCAF is much smaller and simpler. Junior is designed to be added internally to a receiver and use the receiver's audio amplifier and power supply. Both filters feature high performance and simple construction. Experienced builders can assemble either one in a weekend.

The heart of these audio filters is a pair of ICs recently introduced by AMI, the S3528 and S3529. These two ICs can be used together to form an SCF band-pass filter with excellent characteristics. The low-pass and high-pass cutoff frequencies, f_{cl} and f_{ch}, are selected by digital inputs to the ICs at increments of approximately 100 Hz throughout the audio band.

The theory of operation of switched-capacitor filters has been well presented in past issues of *QST* and other amateur and professional electronic journals.[1-4] We will discuss SCF theory only briefly here. Primarily, we will examine the significant features of the S3528 and S3529 and will discuss the construction and use of an audio filter incorporating these devices.

Switched-Capacitor Filters

Whenever an electrical signal is modified in some way (except for pure amplification or attenuation), we say that we have "processed" the signal. Signal processing may be accomplished by continuous or discrete processes. We refer to the continuous process as "analog signal processing" and to the discrete process as "digital signal processing." Examples of analog signal-processing circuits are mixers, detectors, and frequency-selective circuits made from inductors and capacitors. Active filters using op amps, resistors and capacitors also fall into the analog category.

Digital signal processing, on the other hand, relies on a series of "snapshots" or samples of the signal in order to perform a given function. These individual samples are combined and manipulated in a way that yields some desired result. Digital signal processing is used in computerized speech, TV image enhancement and radar. An important part of digital signal processing is digital filtering, which is functionally equivalent to analog filtering. One of several practical digital-filter implementations is the SCF.

The SCF works by storing discrete samples of an analog signal as a charge on a capacitor. This charge is transferred from one capacitor to another down a chain of capacitors forming the filter. The sampling and transfer operations take place at regular intervals under control of a precise frequency source or clock. Filtering is achieved by combining the charges on the different capacitors in specific ratios and by feeding charges back to the prior stages of the capacitor chain. In this way, filters of much higher performance (and complexity) may be synthesized than is practical with analog filters.

The AMI S3528 and S3529

AMI has produced a number of ICs for the telecommunications industry that contain complete SCFs. Two of these circuits, the S3528 and S3529, are of particular interest to the amateur community because of their flexibility and performance. Within the S3528, we find a seventh-order elliptical low-pass filter, a clock generator, a program-

[1] Notes appear at end of article.

mable-clock frequency divider and a pair of buffer amplifiers that are helpful in getting the signal into and out of the IC. The S3529 is similar to the S3528 except that it contains a high-pass filter instead of a low-pass filter. Attenuation is designed to be greater than 51 dB at frequencies above 1.3 f_{cl} for the low-pass filter or below f_{ch}/1.3 for the high-pass filter, where f_{cl} and f_{ch} are the low- and high-pass filter cutoff frequencies, respectively. (In a band-pass configuration, f_{ch} is less than or equal to f_{cl}.) This frequency response characteristic may be seen in the little photo.

A key feature of the S3528 and S3529 pair is the ability to digitally select f_{cl} and f_{ch}. Any of 64 different cutoff frequencies may be selected by setting a 6-bit control code. This code addresses an on-chip ROM whose output controls the frequency divider. In the S3528, the sampling frequency is obtained by dividing the 3.58-MHz clock to equal 40 f_{cl}. In the S3529, the sampling frequency is 44 f_{ch}.

An especially nice set of cutoff frequencies is available in the voice range below 3900 Hz. With a common 3.58-MHz TV color-burst crystal and binary-coded decimal (BCD) inputs, f_{cl} is about 100 times the BCD code on the S3528, and f_{ch} is about 91 times the BCD code on the S3529. Setting the code of both filters to the same value gives a filter whose upper-frequency cutoff is 100 times the switch setting and whose width is 10% of the passband center frequency.

This selection scheme works for all BCD codes between 01 and 39. As you may have observed, there are other digital codes such as 0B and 2E hexadecimal, which lie outside the BCD code set. What happens if you specify one of these codes? You get more frequencies! Some lie between the 100-Hz intervals; others lie outside the 100- to 3900-Hz range, up to 22 kHz. A complete list of codes and frequencies is given in Table 1. Note that codes 35 and 38 deviate from the 100-Hz pattern.

An interesting bit of insight may be gained into the workings of SCFs by examining the possibility of spurious signals in the filter's output. As it happens, there are a few BCD switch combinations that produce very low-level spurious output signals, or "birdies." A few of these artifacts of the digital-filtering process can be heard, although they are much too weak to interfere with communication.

One birdie can be heard when the high-pass switch is set to 00. From Table 1, we see that f_{ch} is 40 Hz. In this case, the S3529 sampling frequency is 1760 Hz. At low-pass switch settings above 16, the tone can be heard. Another can be heard when the high-pass switch is set to 01 and the low-pass switch is set to 39. Here, the sampling frequency of the high-pass filter is 4004 Hz. This is close enough to the low-pass cutoff to get through. Other combinations such as 09/10, 10/11 and 11/12 give rise to weak birdies through the aliasing and quantizing process. An explanation of these signals is beyond the scope of this article.

We were curious about the possibility of the clocks and switched signals causing interference to the station receiver, TVs and so on. Fortunately, we were able to have SuperSCAF tested for emissions at a local facility; it proved to be "clean as a whistle"!

SuperSCAF is a stand-alone unit that contains a switched-capacitor filter, an audio power amplifier and an ac-operated power supply.

Table 1

S3528 and S3529 Cutoff Frequencies

BCD Code	High-Pass Cutoff Frequency (f_{ch}) (Hz)	Low-Pass Cutoff Frequency (f_{cl}) (Hz)	BCD Code	High-Pass Cutoff Frequency (f_{ch}) (Hz)	Low-Pass Cutoff Frequency (f_{cl}) (Hz)
00	40	43	34	3129	3442
01	91	100	35	5423	5965
02	182	200	36	3254	3579
03	273	300	37	3389	3728
04	363	399	38	5801	6382
05	455	500	39	3637	3881
06	546	601			
07	635	699	*Additional Hex Codes*		
08	726	799	0A	433	476
09	822	904	0B	227	250
10	914	1005	0C	904	994
11	1005	1105	0D	935	1028
12	1099	1209	0E	957	1053
13	1179	1297	0F	1043	1147
14	1271	1398	1A	1334	1467
15	1355	1491	1B	1402	1542
16	1453	1598	1C	1585	1721
17	1535	1688	1D	1799	1845
18	1627	1790	1E	1849	2034
19	1731	1904	1F	2034	2237
20	1808	1989	2A	2138	2350
21	1892	2081	2B	2325	2557
22	1985	2183	2C	3687	4057
23	2085	2296	2D	4067	4474
24	2180	2459	2E	4519	4971
25	2260	2486	2F	5085	5593
26	2392	2632	3A	6779	7457
27	2485	2751	3B	8135	8948
28	2543	2797	3C	9038	9943
29	2624	2887	3D	15552	14914
30	2712	2983	3E	16270	17897
31	2805	3086	3F	20338	22372
32	2905	3196			
33	3013	3351			

(courtesy Gould AMI Semiconductors, copyright 1985)

Circuit Description

The block diagrams for the two filters are shown in Fig 1. Both Super and Junior use an identical band-pass filter circuit. Junior's passband is set by binary DIP switches on

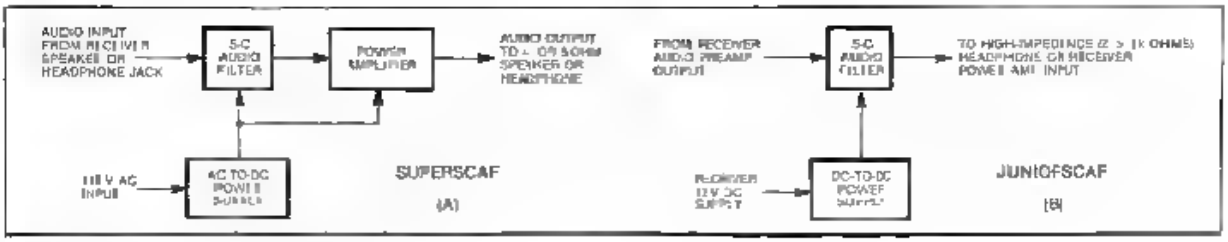


Fig 1—Block diagrams of SuperSCAF (A) and JuniorSCAF (B).

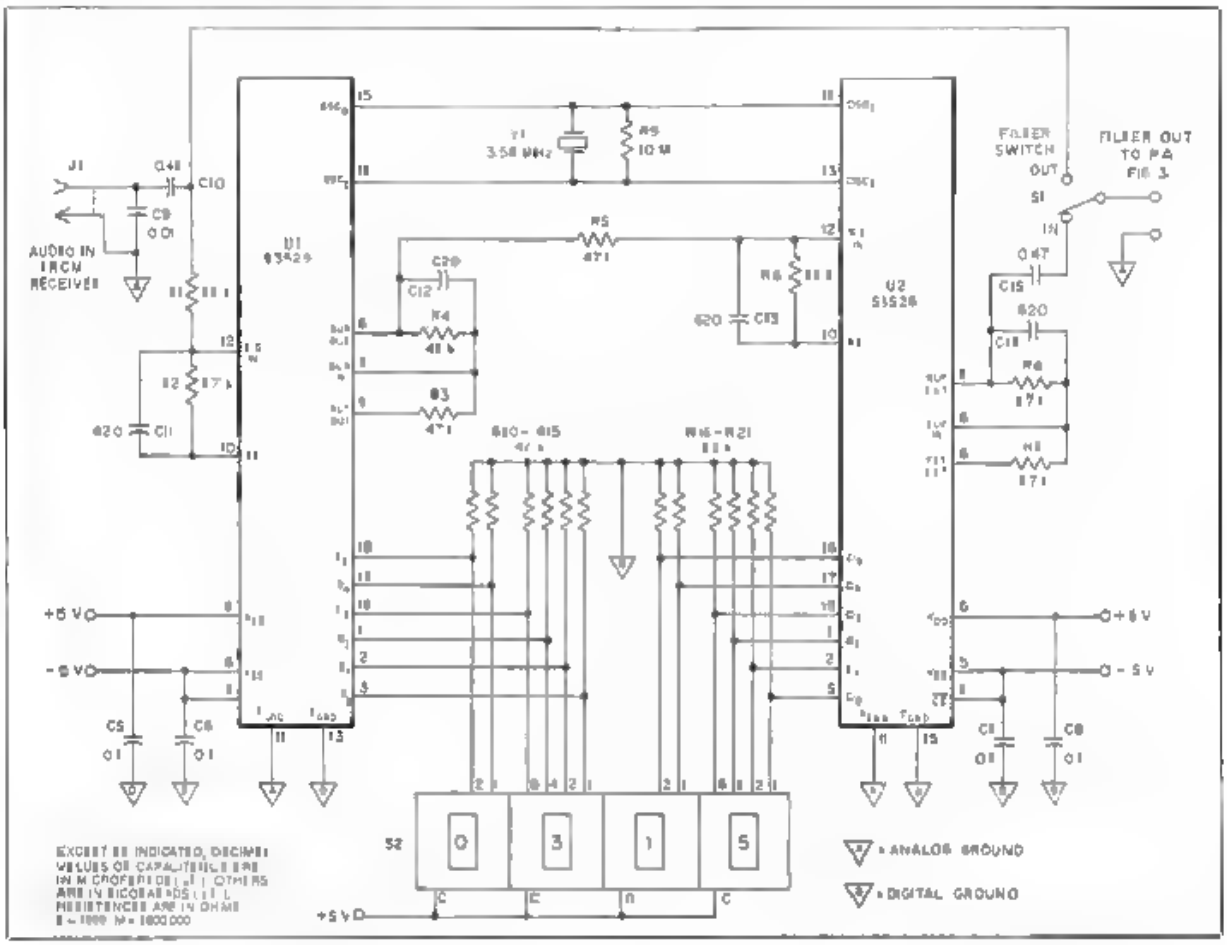


Fig 2—Schematic diagram of the filter sections for SuperSCAF and JuniorSCAF. This circuit, offering simplicity and ease of construction is used for both filter units.

J1—Phono jacks.
J3—¼-inch phone jack.
R10-R15, R16-R21—47 kΩ × 7 resistor packs used (one resistor in each pack not connected).
S1—SPDT toggle.
S2—BCD thumbwheel switch (SuperSCAF), C & K 342710640-01904 or equiv. For JuniorSCAF, two 8-position DIP PC-mount switches are used.
U1—AMI S3529 programmable high-pass filter.
U2—AMI S3528 programmable low-pass filter.
Y1—3.58-MHz TV color-burst crystal.

the PC board. SuperSCAF's passband is controlled by thumbwheel switches on the unit's front panel. Super also has its own audio power amplifier and ac-operated power supply. In reading the following circuit descriptions, keep in mind that SuperSCAF is a self-contained unit that accepts low- or high-impedance inputs and delivers 1.5 W of audio output at 8 ohms. Junior, on the other hand, has a high-impedance output circuit. It can drive high-impedance phones directly, but doesn't have the "oomph" to drive a speaker.

Refer to Figs 2-4. The input signal to SuperSCAF is obtained directly from the speaker output or the headphone jack of your receiver. The signal is passed first into the S3529 high-pass filter and then into the S3528 low-pass filter. A pair of switches sets the frequency of each filter. The filters are followed by an audio power amplifier. Switching is provided to bypass the filter if desired.

As with any digital filter, it is necessary to band limit the input signal to prevent aliasing. The combination of receiver IF-stage filters and a bit of high-frequency rolloff in the audio sections of most receivers is sufficient to prevent problems. C11 and R2 are used in conjunction with

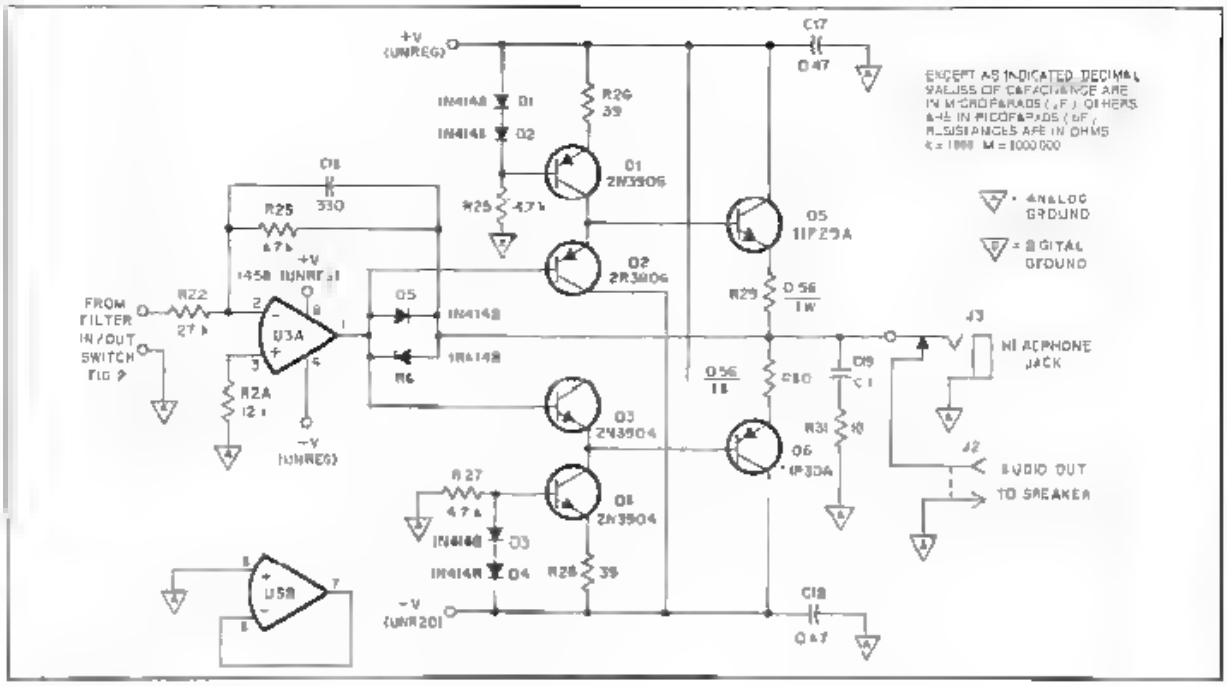


Fig 3—The SuperSCAF audio amplifier schematic diagram.
D1-D6, incl—1N4148 silicon diode.
J2—Phono jack.
Q1, Q2—2N3906.
Q3, Q4—2N3904.
Q5—TIP29A.
Q6—TIP30A.
U3—MC1458 dual op amp.

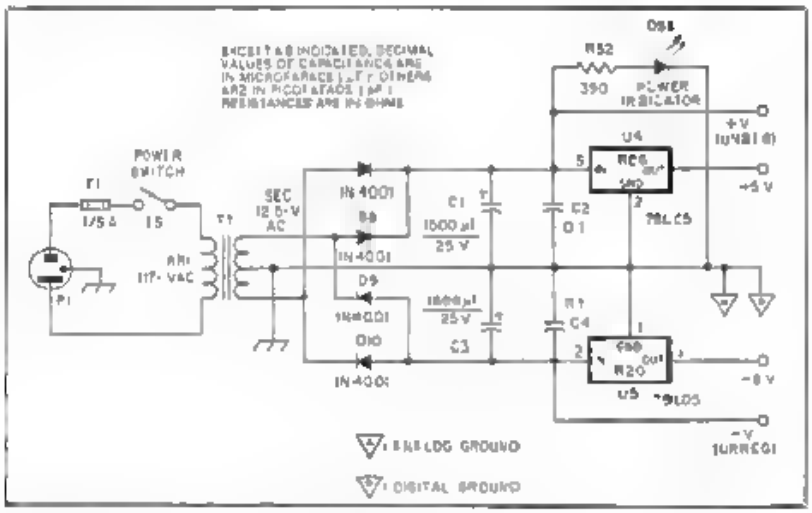


Fig 4—Schematic diagram of the power supply designed for use with the SuperSCAF.
D7-D10, incl—1N1001.
DS1—LED.
S3—SPST toggle.
T1—117-V pri; 12.6-V sec, 1A (RS 273-1505 or equiv).
U4—7805 5-V positive regulator.
U5—7905 5-V negative regulator.

the input op amp of the S3529 to form a simple analog low-pass filter, just in case. The six frequency-select lines to each IC are pulled to digital ground by 47-kΩ resistors, representing a logic low. The BCD switches then selectively apply +5 V to the lines, depending on the code, to indicate a logic high.

Both ICs share a common 3.58-MHz crystal and 10-MΩ resistor. In addition to economy, this scheme ensures that both filter ICs operate synchronously from the same clock. The output signal of the S3529 is smoothed by the analog low pass filter made up of R4, C12 and the output buffer. An additional stage of analog filtering is provided by R6, C13 and the input buffer of the S3528.

The low-pass filter functions similarly. Output from the low-pass filter is smoothed by the S3528 output op amp, R8 and C14. The filtered signal is then passed to the power amplifier.

Although monolithic audio amplifier ICs are readily available, a discrete-component power amplifier (Fig 3) was designed for SuperSCAF. (This choice was dictated by the split power supply discussed later.) The power amplifier is basically a voltage amplifier composed of U3 followed by a current amplifier. Q2 and Q3 act as drivers for the output transistors, Q5 and Q6. Q1 and Q4 act as constant-current sources for the driver collector and output transistor base nodes. Short-circuit protection (SCP) is provided by the current-limiting action of D5 and D6. The power amplifier will deliver a maximum of 1.5 W to a 4- or 8-ohm load, more than enough for a comfortable listening level. Trying to drive the amplifier beyond 1.5 W output will result in distortion.

A split power supply (Fig 4) is used to simplify the input and output signal-return path and to accommodate the ±5-V supply requirements of the S3528 and S3529. Supply voltages for the S3528 and S3529 are obtained from a pair of low-current complementary regulators. Separate analog and digital grounds are used to prevent digital noise from appearing on the analog ground return. The two ground systems are joined at the power supply.

JuniorSCAF (see Figs 2 and 5) is ideal for QRP work. Since the power amplifier

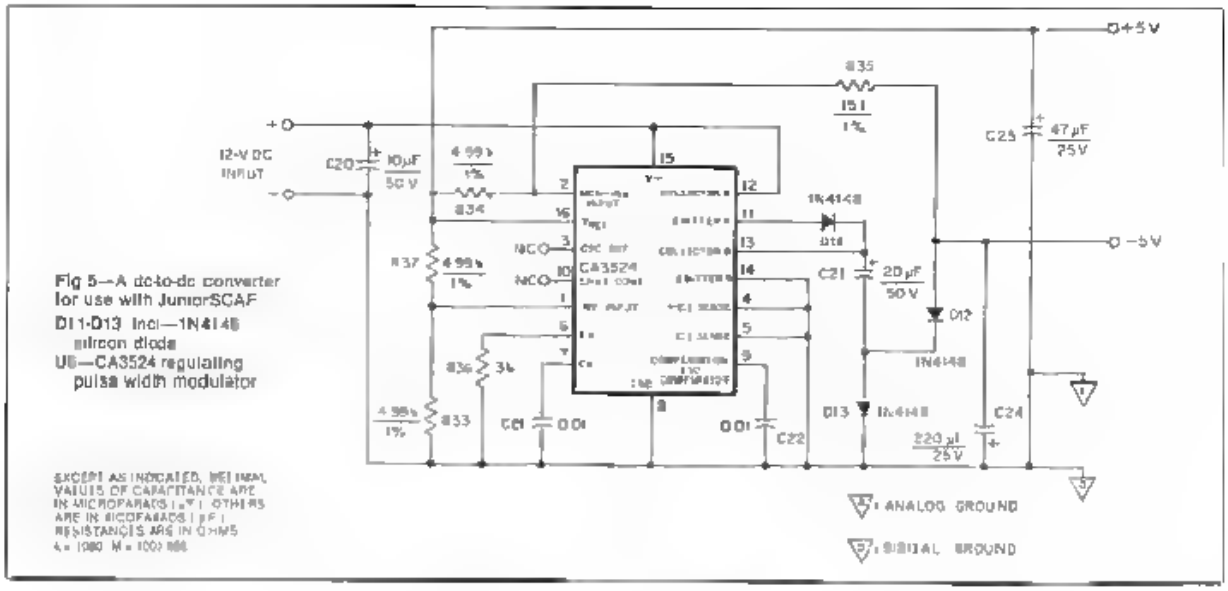
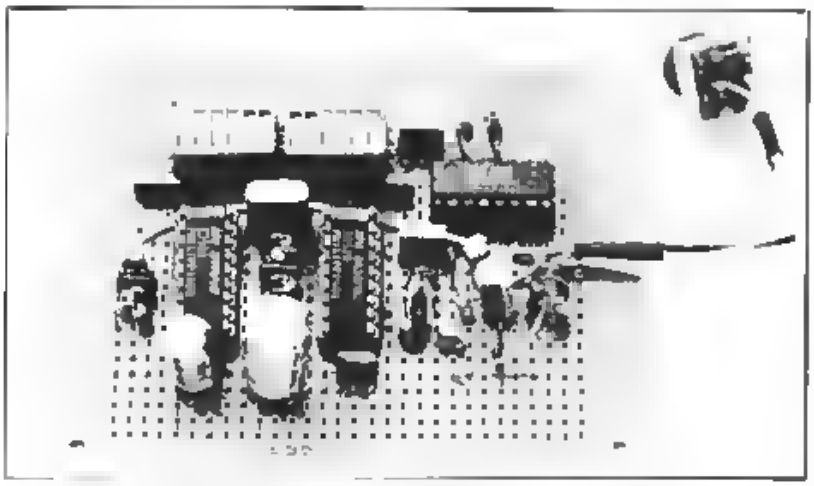


Fig 5—A dc-to-dc converter for use with JuniorSCAF.
D11-D13 incl—1N4148 silicon diode.
U8—CA3524 regulating pulse width modulator.

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF); OTHERS ARE IN PICOFARADS (pF). RESISTANCES ARE IN OHMS: k = 1000, M = 1 000 000.



JuniorSCAF is designed to add a host receiver's 12-V ac supply and audio power amplifier

accounts for most of the operating current, its elimination allows the two complementary 5-V dc power supplies to be derived from a simple dc-to-dc converter operating from a 12-V dc source within the receiver.

Although JuniorSCAF is the simpler of the two filters, to install it in a receiver it is necessary to break the signal path between the receiver's audio preamplifier and power amplifier. The output from the preamp is coupled to Junior's input. Junior's output is connected to the receiver's audio power amplifier input. Also, it's necessary to tap into a well-filtered supply of between +12 to 40-V dc to obtain operating power. Because these details vary widely from receiver to receiver, we can't offer more specific installation instructions. Unless you are comfortable cutting leads and traces inside your equipment (or can find a friend to do it for you), we suggest you build SuperSCAF instead.

Construction

Assembling these filters is straightforward. Although the layout is not critical, it's always best to keep leads as short as possible. If you decide to use perf board instead of the PC board, be aware that the pinout of the S3528 is slightly different from that of the S3529.

An interior view of the SuperSCAF is shown in Fig 6. A metal box is used as an enclosure for the prototype. Metal is preferred to plastic because of its strength and also because it offers a degree of RFI protection. Remember that the filter may be required to work in an area of high RF-signal strength.

The rectangular hole for the BCD switches is cut with a nibbling tool. Drill a pilot hole large enough to accommodate the nibbler in the center of the BCD switch mounting location. Next, the sides of the switch hole are cut by the nibbler. Finally, the edges of the hole are filed until smooth. Although we used several types of BCD switches during the course of the project, the one we like best is made by C & K components (see parts list). This switch is small, but has a smooth feel and clearly legible digit markings. The high- and low-pass switch positions have stops installed that limit the range to between 00 and 39, matching the filter's operating range.

After holes for the other switches, jacks, power cord and LED are drilled, these components are mounted and connected to the circuit board. We like the looks of a small (1/8-inch diam) LED ac-power indicator. A hole for the LED must be drilled for a snug fit. The LED is then held firmly in place by a drop of glue on the back. Color-coded ribbon cable works nicely for attaching the switches and LED to the circuit board.

The circuit board is mounted to the bottom of the case by stand-offs. To dissipate the heat and prevent thermal runaway, the output transistors *must* be mounted to heat sinks. We used the rear of the case as a heat sink (see Fig 6). The output transistors must be insulated from the chassis by mica washers and an insulating screw washer to prevent short-circuiting the supply voltages. Use thermally conducting silicon grease on both sides of the mica washers.

For safety reasons, a 3-wire power cord should be used. Connect the ground conductor (green) to the chassis and connect the neutral conductor (white) directly to the

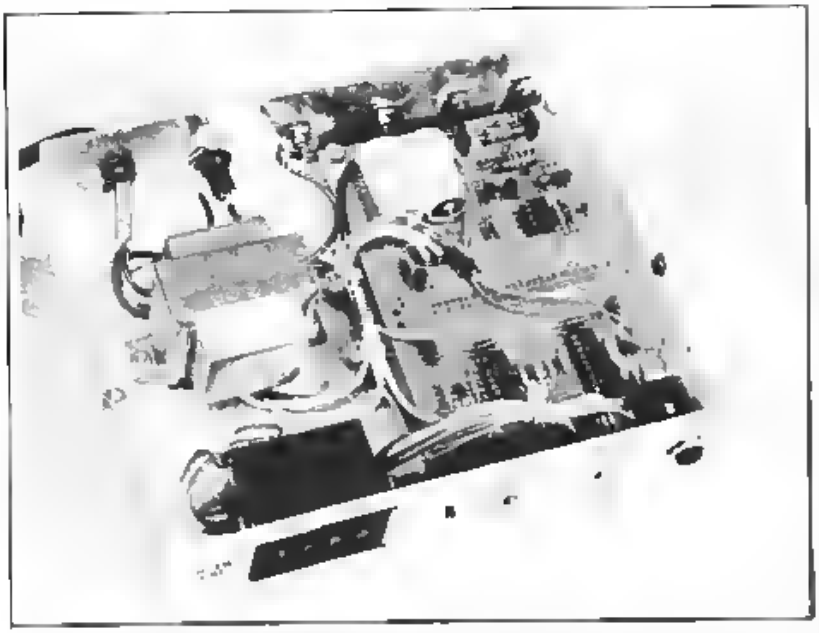


Fig 6—An interior view of SuperSCAF. Note the mounting of the audio-amplifier output transistors, the rear panel of the enclosure is used as a heat sink. The enclosure used for SuperSCAF is a Radio Shack part (RS 270-272).

primary of the power transformer. Solder the hot (black) wire to the spring contact at the rear of the fuse body. Connect the sleeve terminal of the fuse to the power switch. Be sure that the power connection is wired in this manner. Failure to do so may result in a serious shock hazard.

Performance and Operation

Connect the receiver speaker output to the AUDIO IN jack. Plug the speaker into the AUDIO OUT jack. Use shielded audio cable to reduce the possibility of introducing RFI into the filter.

The SuperSCAF and Junior are a pleasure to use. If you mate them with an older rig and operate CW, you'll be surprised by the sudden quiet in the shack. Under many conditions, noise and QRM simply disappear. We became aware of ac hum in one of our receivers only after SuperSCAF made it go away! The filter even does a respectable job on the woodpecker and "sons of the woodpecker." There is no artificial ringing, only the residual noise within the filter passband.

The effect of the filter on SSB signals is not as dramatic, but certainly noticeable and worthwhile. Simply set the switches to 03/27 and eliminate splash outside that frequency range. When conditions get tough, experiment with a narrower passband. Setting the low side of the passband below 02 is never needed and is an open invitation to aliasing.

The most significant operation difference between SuperSCAF and Junior is the passband switching. If Junior is mounted inside a receiver, it is inconvenient to change the passband during operation. We recommend that Junior be set up for a passband of about 500 Hz for CW and switch settings of 03/24 (300-2400 Hz) for SSB.

An obvious advantage of the thumbwheel switching scheme is direct passband readout. Another is the ability to adjust the upper and lower cutoff points in small steps, hearing the effect as you go. For narrow-band interference such as "tuner-uppers," the interference will often disappear at a particular step. At 24, you hear him, at 23, he's gone. For wideband interference, the effect is not as dramatic.

On CW, RTTY and other narrow-band modes, the filter performance is spectacular (see Fig 7). We both work a lot of CW and have older rigs with SSB crystal filters having passbands that are much too wide for comfortable code reception. With SuperSCAF, we get tremendously improved selectivity.

Your new-found selectivity requires changes in operating habits. If the filter is set so that the passband is narrow, say 07/07 (about 70 Hz wide), the band may seem empty. The problem is that your accustomed tuning rate may be too fast for such a narrow bandwidth. You may tune completely across a station during the time between code elements and never hear the

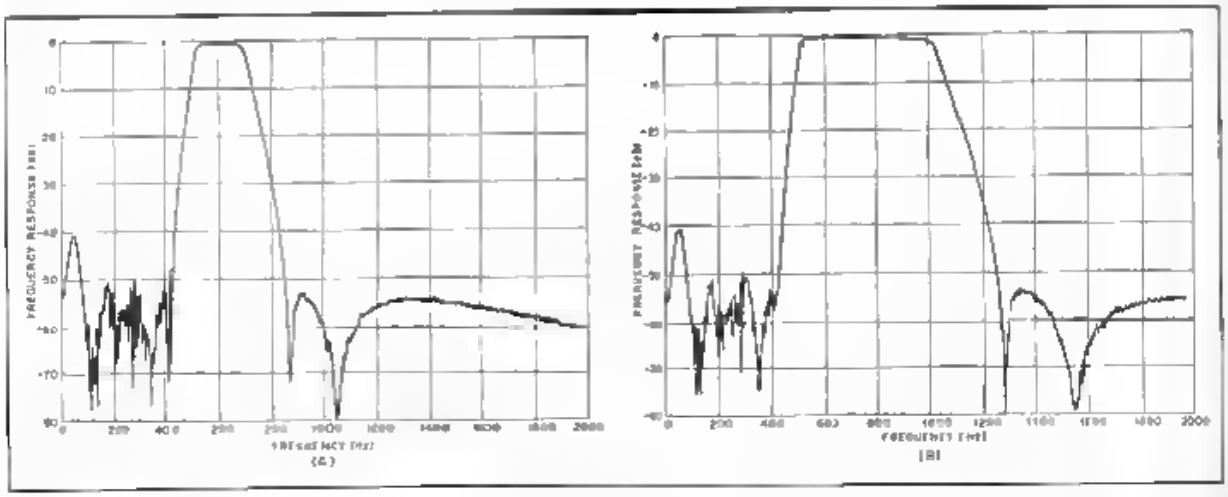


Fig 7—Spectral plots of the SuperSCAF response made in the ARRL lab. Note the steep filter skirts. For both plots, the center frequency is 1 kHz; vertical divisions are each 10 dB and horizontal divisions are each 200 Hz. At A, the front-panel switches are set to 05/06; at B, 05/09. Those settings equate to passband widths of about 200 and 500 Hz, respectively.

signal. The solution is to search the band using a relatively wide passband or with the filter bypassed. When you find a "live one," close the passband around him. We often use a setting of 05/09 for search, and then narrow the passband to 07/07 for the QSO.

Be alert to frequency drift, particularly when you turn things over to the other station. It's very easy for one of you to slip outside a 70-Hz passband. If the other station is not where *you* expect it, widen the filter passband to re-acquire the signal, then narrow the passband on the new frequency. Also, experiment with disabling the AGC if your receiver allows that. Sometimes a strong signal within the IF passband will grab your AGC and reduce the incoming signal levels to practically nothing. You might not hear the interfering station, but you'll know it's there.

Summary

The possibilities presented by monolithic SCFs are numerous. We have built several variations on the theme presented here, and all have worked well. One unit was powered by a pair of 9-V batteries and used an IC power amplifier instead of the discrete amplifier of Fig 3. Another unit included a tone decoder to supply a digital signal to a computer for receiving Morse code and RTTY. That unit was mounted in the transceiver's companion speaker box.

At the outset, our goal was to design an easily constructed audio filter with excellent performance. We are pleased with the results in every way—we hope you will be, too.

Notes

[1]R. Schellenbach and F. Noble, "Switched-Capacitor Filters—An Emerging Technology for Amateur Radio Use," *QST*, Mar 1984

[2]R. Olsen, "Digital Signal Processing for the Experimenter," *QST*, Nov 1984

[3]AMI Telecommunications Design Manual, Gould AMI, 1982 3800 Homestead Rd, Santa Clara, CA 95051

[4]J. Connel, "Switched-Cap Filters Mate With Microprocessors," *Electronic Products Magazine*, Sep 3, 1984

[5]We wish to thank Don Fisher W4PLA, and the NCR Corporation Emission Testing Service, Lake Mary, Florida, for providing the FCC Part 15, Class B test data for the SuperSCAF audio filter.

[6]Gould Semiconductors sells the S3528 and S3529 ICs through a network of distributors. Call Gould at 208-233-4690 for the name of their nearest distributor. New Horizons Electronics Corporation, 6000 New Horizons Blvd, Amityville, NY 11701, 516-226-6000, will sell to individuals (prepaid) with a $25 minimum order. (The chips cost about $7 each at the time of publication.)

Ahronics sells kits and completed SuperSCAFs, but not parts. Check the *QST* Index of advertisers for their listing, which contains current sales information.—*Ed.*

From July 1986 QST, p 34.

The SWR Twins—QRP and QRO

Part 9: Portable amateur operation often calls for miniature equipment. Here are two tiny SWR indicators—one for QRP and one for high power.

By Doug DeMaw, W1FB
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Does the inconvenience of too-large SWR-indicating gear complicate your portable operations? It is not uncommon for us to feel that some of the commercially made SWR bridges and RF-power meters are too big and too costly for occational use during field day, camping trips, vacations and even DXpeditions. I have seen SWR meters that were larger than an entire QRP station, which presents a rather absurd picture! Because of my need for small accessory equipment, I have built a number of compact Transmatches and SWR meters. The unit we shall consider in this article was built to provide an example of small units that you can build inexpensively for field use. We will also consider some practical ideas for home construction that can be applied to other projects as well. These SWR indicators are not works of art, at least from an aesthetic point of view, but you can easily impart a professional appearance to them if you are skilled in the craft of cabinet and panel design.

Do You Need an SWR Indicator?

SWR meters and RF-power indicators have become a way of life with most of us. But, "way back when," we managed quite well without these sophisticated gadgets. An experienced amateur could tell if the antenna SWR was low by observing the settings of the tune and load controls of the transmitter. That is, the plate tuning and loading controls were at approximately the same settings as when the transmitter was connected to a dummy load of the appropriate impedance, thereby indicating a low SWR. Some of us used RF ammeters in the feed line to indicate maximum RF current, a condition that generally occurred when the feed line was matched to the antenna feed point.

SWR has become a more significant concern today because of the many solid-state transmitters that exist. They must "look" into a low SWR—usually 2:1 or less—in order to develop the rated output power and to protect the final-amplifier transistors from damage. The built-in SWR-protection circuits reduce the transmitter output power as the SWR increases. Therefore, it is helpful to have an SWR indicator between the transmitter and the transmission line. The antenna can then be adjusted by means of its length or matching circuit to obtain a low SWR reading.

SWR indicators are useful also as relative output-power meters. They help us to keep tabs on the antenna system and the transmitter performance. Most SWR instruments can be calibrated to read RF power as well, and we will discuss this principle later in the article.

A QRP SWR/Power Meter

Neither of the instruments in this article is new in concept. The resistive QRP bridge was developed many years ago by the late George Grammer, W1DF. The QRO bridge is a design product of Warren Bruene of Collins Radio. The latter design has become the standard for most amateur SWR and power meters of commercial origin. A number of variations in the basic designs have been introduced, along with some extra convenience features.

Fig 1 shows the circuit for our low-power SWR bridge/RF power meter. R1, R2, R3 and R4 comprise a 50-ohm dummy load. Some of the RF voltage developed across the load is sampled through R5 and supplied to the resistive bridge that consists of R6, R7 and R8. The antenna represents the remaining leg of the bridge. When it reflects a 50-ohm condition, the bridge is balanced and the meter reading falls to zero. D1 rectifies the RF voltage to provide dc for the metering circuit. Additional examples of this general circuit are given in *Solid State Design for the Radio Amateur* (temporarily out of print).

R10 is a panel control that is used to establish the "sensitivity" of meter response versus the power level. R11 is a PC-mounted potentiometer that we can use to calibrate the meter for a full-scale reading of 10 W. Once set, it should need no further adjustment.

Since R1, R2, R3 and R4 have a combined rating of 8 W, we must not permit a sustained RF power amount of more than 4 W to be fed into the instrument, lest the resistors become damaged from excessive heating. Momentary tests with powers up to 10 W will not harm the resistors, provided the key-down period does not exceed 15 seconds. Allow a cool-down period of at least 30 seconds between brief tests with more than 4 W of RF power. Film resistors are used in my model, but 5%, 2-W carbon-composition resistors will work equally well. If you cannot locate them, you may purchase the film resistors by mail.[1]

The power handling capability of this instrument may be increased by using higher-wattage (noninductive) load resistors or by connecting an external dummy load to replace the built-in one. *Warning*: If you plan to use more than 10 W of RF power, and a larger dummy load, be sure to in-

[1]Notes appear at end of article.

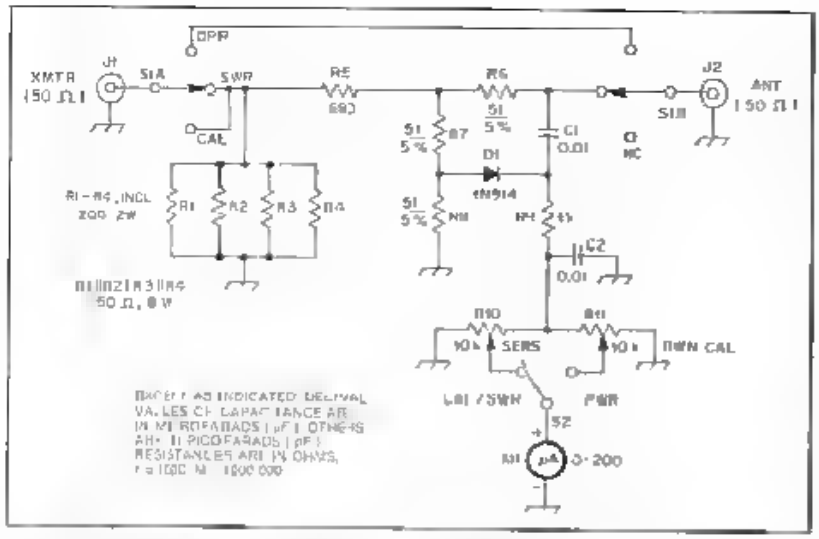


Fig 1—Schematic diagram of the QRP SWR bridge. Resistors are carbon-composition types. Capacitors are disk ceramic. Part numbers listed below are Radio Shack designators except when otherwise noted.

C1, C2—Disc ceramic, RS 272-131.
D1—Small-signal silicon diode, RS 276-1122.
J1, J2—RCA style single-hole-mount phono jack, RS 274-346.
M1—Miniature microammeter, 0-50, 0-100 or 0-200 μA. See note 2.
R1-R4, incl—200-ohm, 2-W noninductive resistor. See note 1.
R5—680-ohm ½-W resistor, RS 271-021.
R6, R7, R8—51-ohm, ½-W noninductive resistor. Available from All-Electronics Corp, Los Angeles, CA. Radio Shack 47-ohm units (271-009) may be substituted for suitable performance.
R9—1-kΩ, ½-W resistor, RS 271-023.
R10—Panel-mount control, 10-kΩ, linear taper carbon-composition, RS 271-1721. Knob (0.5-inch OD) is RS 274-403.
R11—Trimmer control, PC mount, 10-kΩ, RS 271-335.
S1—Two-pole, three-position rotary wafer, RS 275-1386 (three positions not used).
S2—SPDT miniature toggle, RS 275-613.

Fig 2—Meter scale that may be pasted over the original scale of the meter cited in note 3. See text for method of making your own custom scale at ×4.

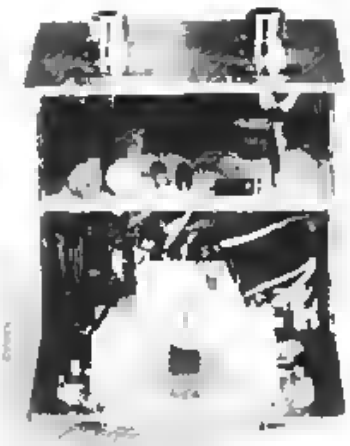


Fig 3—An interior view of the QRP SWR meter. The PC board is mounted vertically by means of two no. 6 spade bolts. A ¼-inch hole is drilled in the rear panel of the box to provide access to the RF-power calibration control, R11.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

create the value of R5 to prevent excessive RF current from flowing in the bridge circuit. Sample only enough RF energy to provide a full-scale meter indication (R10 set for maximum sensitivity) at about half the power level you anticipate. In other words, if you expect to use 50 W of RF power, select an R5 value that will give a full-scale meter reading at 25 W.

How to Use the QRP Meter

Calibration of this instrument was covered in Aug 1983 *QST*, at which time I described a similar instrument.[1] I will review the operating procedure here, since some of you may not have used this type of bridge for SWR and power measurements.

S1 allows us to bypass the bridge after making SWR or power measurements. The bridge is out of the circuit when S1 is in the OPR position. When we switch to the CAL mode, the bridge has no terminal im-

This enables us to adjust R10 (RF power applied) for a full-scale reading at M1. Next, we move S1 to the SWR position. The meter then indicates the relative reflected power. If the antenna is matched and tuned properly for a 50-ohm condition, the meter will read zero. If not, the antenna system or Transmatch should be adjusted until the meter reads zero. Once this is achieved, set S1 to the OPR mode.

RF-power measurements may be made (after M1 has been calibrated by means of R11) by placing S1 in the CAL position and setting S2 to read PWR. This removes the antenna (12) from the circuit and permits us to develop RF voltage across R1, R2, R3 and R4. You may feed various power levels (from 1 to 10 W) into the circuit, then note the meter reading for each power amount. A calibration scale may then be drafted for future reference. The 1-10 numerical scales on the meters of these SWR indicators were drawn by hand at ×4. I used press-on decals for the numbers. I then had the meter scale reduced ×4 at a "quick-print" shop, at a cost of 24 cents each. The new scales were pasted over the original faces of surplus 200-μA S meters.[2] You may use an available meter that has a sensitivity of 50 or 100 μA. Fig 2 contains a 0-10 meter scale that you may cut out or photocopy for use on the meters that are available from the source listed in note 3. The cases come off easily, and the meter face can be popped out for modification.

The interior of the QRP bridge is shown in Fig 3. A scale parts-placement guide for the PC board is provided in Fig 4A.

QRO SWR Indicator

This fraternal twin to the QRP bridge will measure SWR and RF power at levels up to 1 kW. The major problem is that the instrument is so tiny and lightweight that the coaxial attachment cables may become the "tail that wagged the dog." This is often a penalty associated with miniature gear. I find that RG-8X 50-ohm cable minimizes the problem. I have experienced frustration when trying to use the heavier, stiffer RG-8 cable.

Fig 5 shows the circuit for the QRO bridge. I used a hybrid diagram in order to clarify the relationship of T1 to the rest of the circuit. T1 is a transformer for sampling RF current in the feed line. The cable that passes through the toroid core serves as a one-turn primary winding for T1. QRP versions of this bridge can be built if we use a two-turn link in place of the straight conductor that passes through the toroid. This will increase the sensitivity.

The bridge (minus the cabinet) is suitable for inclusion in Transmatches. The PC board (Fig 4B) can be installed near the RF-input jack of the Transmatch. The leads that go to S1 are not critical as to length, so S1, R1 and M1 may be panel-mounted in your Transmatch, if desired.

The shield braid of the pass-through coaxial line (T1) is grounded only at one end. This provides a Faraday shield to discourage the flow of harmonic currents into the bridge. C1, C2 and C5 form a capacitive divider for balancing the bridge in a 50-ohm circuit. D1 and D2 provide for the metering circuit. Germanium diodes are

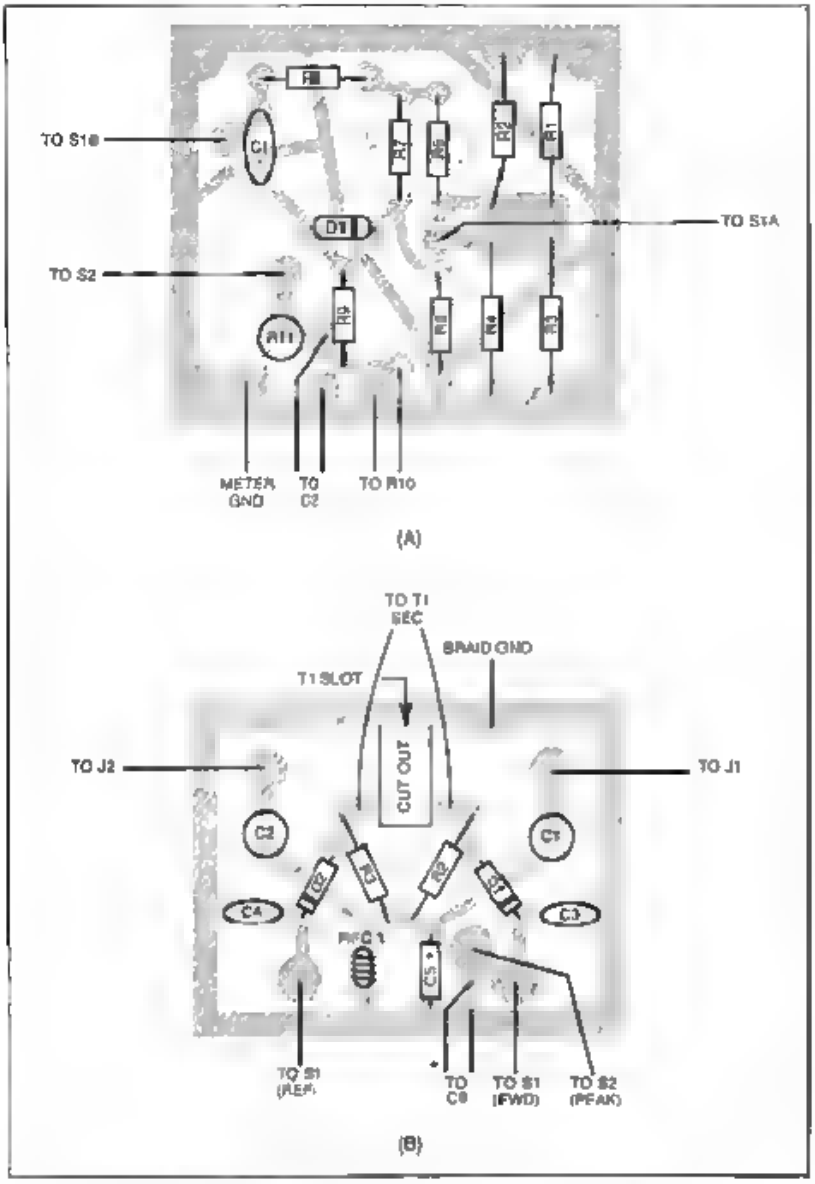


Fig 4—Scale parts-placement guide for the QRP meter (A) and the QRO meter (B), as viewed from the component side of the boards.

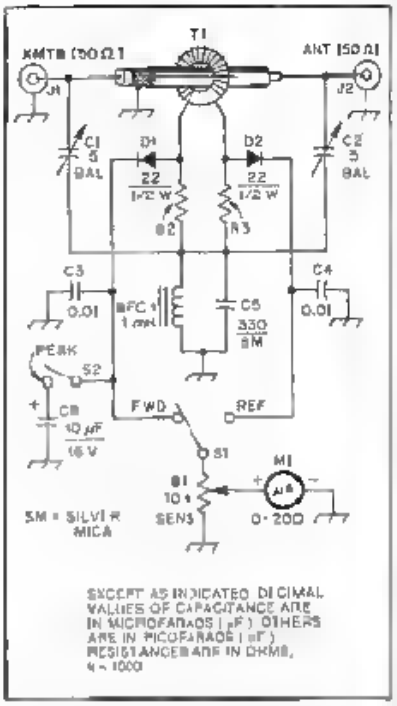


Fig 5—Hybrid diagram of the QRO SWR bridge. A short length of 50 ohm coaxial cable is passed through the center of toroid T1, as indicated. Fixed-value resistors are ½-W carbon-composition types. Other components are described below. Radio Shack numbers included.

C1, C2—Miniature 1-5 or 1-8 pF air or piston trimmer. See note 4.
C3 C4—Disc ceramic. RS 272-131.
C5—Silver mica or polystyrene, 330 pF. NP0 ceramic also suitable. Silver-mica capacitor avail. from All-Electronics no. DMCP-330.
C6—Tantalum or electrolytic. RS 272-1436.
D1, D2—Small-signal germanium diode, RS 276-1123, 1N34A.
J1, J2—Single-hole mount BNC or connector of your choice, RS 278-105.
M1—Miniature microammeter, 0-50, 0-100 or 0-200 µA. See note 2.
R1—Linear-taper, carbon-composition, panel-mount control, 10 kΩ, RS 271-1721.
R2, R3—22-ohm, ½-W carbon composition, RS 271-005.
RFC1—Miniature RF choke, 1 mH. Avail. from All-Electronics Corp, no. CC-1000 or from RCD Electro.
S1—Miniature DPDT toggle, RS 275-613.
S2—Miniature SPST toggle. RS 275-612.
T1—60 turns of no. 30 enam. wire on an Amidon Assoc. T50-2 powdered-iron toroid core. Mount in slot on PC board (see text).

recommended to ensure sufficient dc voltage when small amounts of reflected power are present. Silicon diodes, such as 1N914s, may cease to conduct (0.7 V approx) before the SWR is adjusted for a 1:1 state, thereby causing us to think we have a match when, in fact, we do not. Germanium diodes have a lower barrier voltage (approx 0.4). This isn't a problem with the circuit of Fig 1.

I used RG-8X for the line that passes through T1, mainly because it provides a close fit to the center hole of the toroid core. Or, you may use the smaller RG-58, then affix it in the toroid with silicone cement. The RG-8X has a higher power rating, however.

C6 is included with S2 to provide a leveling effect of the meter reading during SSB operation. It will enable you to get an approximate peak-power reading if you calibrate this instrument to read RF watts.

Meter calibration (watts) can be accomplished if we feed a known amount of power through the bridge (into a 50-ohm noninductive load) and adjust R1 for a full-scale reading at M1. A panel mark is then made for this setting of R1. It will enable us to readjust R1 later on for reading RF power. Once we identify this setting of R1, the meter scale can be plotted at different power levels, as I suggested for calibrating the QRP bridge of Fig 1. An RF probe, VTVM or FETVOM and a 50-ohm load may be used for calculating the transmitter output power by measuring the RF voltage across the 50-ohm load [P(watts) equals $V(RMS)^2/R(ohms)$].

Adjustment of the QRO bridge is done with a 50-ohm dummy load connected to J2 of Fig 5. Apply RF power with S1 in the FWD position. Adjust R1 to provide a full-scale reading at M1. Switch S1 to the REF

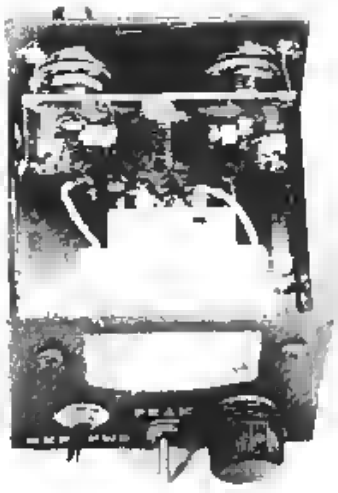


Fig 6—Interior view of the QRO SWR bridge. The PC board is attached to the solder terminals of the BNC jacks. The bottom edge of the PC board is soldered to the chassis at the center. Short wires (upper right and left of PC board) ground the board to the rear panel of the box. C1 and C2 are on the etched-foil side of the board to permit easy access during bridge adjustment (nulling).

mode and observe the meter reading. If it is not zero, adjust C2 for a zero reading. Next, reverse the cables at J1 and J2 and interior view of this bridge is provided in Fig 6.

Construction Notes

The cabinets for these units are made from PC-board pieces. The box dimensions (hwd) are 2 1/2 × 2-5/8 × 3 inches. I chose case-metal aluminum sheeting for the box covers since it was available at the hardware store. This is an advantage for the QRP bridge, since the holes in the cover permit air flow around the load resistors.

My cabinets were formed by soldering together sections of double-sided PC board (front, back and bottom plates). Strips of PC board are used as stabilizing members between the front and rear panels, adjacent to the bottom plate. These strips provide anchor points for the top cover, which is affixed by means of no. 6 sheet-metal screws. I cut the meter holes with a hand-operated nibbling tool.

I discovered by chance that Krylon[3] grey undercoat spray paint is excellent on panels. It was the only can of paint I had on hand when I built these units, so I used it. Not only does it dry quickly (5 minutes), it provides a tough matte finish that is quite immune to smudging from our fingers. It appears to be an excellent paint for amateur projects. If you prefer a gloss finish, you apply RF power. Set C1 for an M1 reading of zero with S1 set in the FWD position. Repeat this procedure one more time. C1 and C2 may be any small trimmer of quality, such as miniature air variable or glass piston trimmers.[4] The *minimum* capacitance of the trimmer must be 1 pF or less in order to null the bridge. An

may spray the grey panels with polyurethane varnish (also available in spray cans).

The front panels of my units look a bit crude because of the black Dymo® tape labels. Grey Dymo tape would provide a much nicer appearance. White press on decals might be an even better choice for the control labels.

Adhesive backed plastic feet are affixed to the bottom of the boxes to prevent excessive movement of the bridges and to avoid scratching the surfaces of desks or tables on which they rest. Screw-on feet may be substituted.

Either of the bridges can be made smaller, should that be your pleasure. I allowed substantial wasted space in order to keep the units in a size class that would not be awkward to work with (the "tail that wagged the dog" problem).

I hope you have fun with one or both of these weekend projects. You should enjoy building these bridges, and they will not endanger your project fund significantly!

Notes

[1]Deleted

[2]D. DeMaw, "A Beginner's Look at RF Power Measurement," *QST*, Aug 1983, p 35.

[3]Most edgewise mounted audio or S meters have microampere movements. Meters used in the instruments described here are available from the supplier in note 4.

[4]Piston trimmers suitable for this project are listed in the BCD Electric catalog, PO Box 830119, Richardson, TX 75083-0119.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

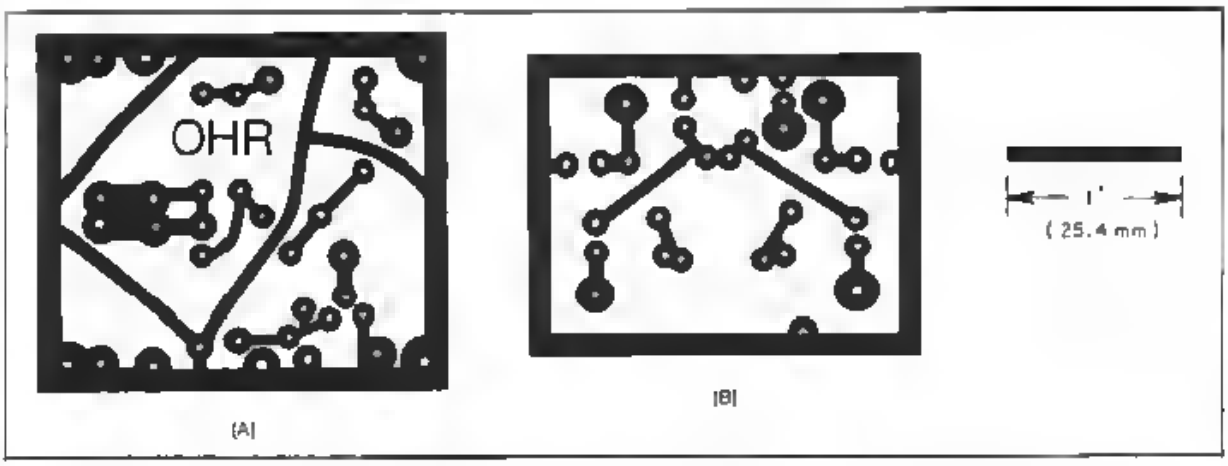


Fig 7—Circuit-board etching patterns for the QRP (A) and QRO (B) SWR/power meters. The patterns are shown full-size from the foil side of the board. Black areas represent unetched copper foil.

A NEW FACE FOR A RECALIBRATED METER

☐ In "A Simple and Accurate QRP Directional Wattmeter," (pp 19-23 and 36, this issue), I described a QRP wattmeter that uses a standard 0-1 milliammeter modified with a custom, nonlinear scale calibrated directly in power and SWR according to the values shown in Table 1. Making new scales for a stock meter is one solution; adding markings to the existing meter scale is another. It's fairly easy to make the new scales readable, and somewhat harder to make them look nice. If you decide to make new scales rather than add marks to the existing scale, you'll want to record the correct places to make the new marks before you obliterate the old scale. One way to do this is as follows.

Refer to Fig 1. Attach the meter face to a large piece of paper. Trace around the face so you can exactly reposition it later. Find the meter pivot point by extending the tick marks at the scale's ends, and verify this point by extending a couple of other points on the scale. Then draw lines from the pivot point through the meter face to an area beyond the face, labeling them appropriately, as shown in Fig 1. Then you can remove, repaint and replace the meter face. The new scale and marks can be hand-drawn; press-on letters and numbers can be used, or the face may be made with a photographic process.

A caution: Anyone who sees my home-built equipment immediately realizes that, although I'm willing to spend a lot of time on functionality and performance, I don't devote much time to beauty! So you'll have to look elsewhere for advice on how to make a meter face good-looking. When finished, the meter face should resemble Fig 2.—*Roy Lewallen, W7EL, 5470 SW 152 Ave, Beaverton, OR 97007.*

Table 1

Meter-Recalibration Data for the QRP Wattmeter

Power	*Meter*	*SWR*	*Meter*
0.0	0 0	1	0 0
0 05	0 224	1 5	0.200
0 1	0 316	2	0 333
0 2	0 447	3	0 500
0 3	0 548	4	0 600
0 4	0 633	5	0 667
0 5	0 707	7	0 750
0 6	0 775	10	0.818
0 7	0 837	∞	1 0
0 8	0 894		
0 9	0 949		
1 0	1 0		

The *Meter* column expresses fractions of full-scale readings on the original meter scale. For example, the new SWR = 3 mark should be placed at the same place as the half-scale (0 5) mark on the original meter face.

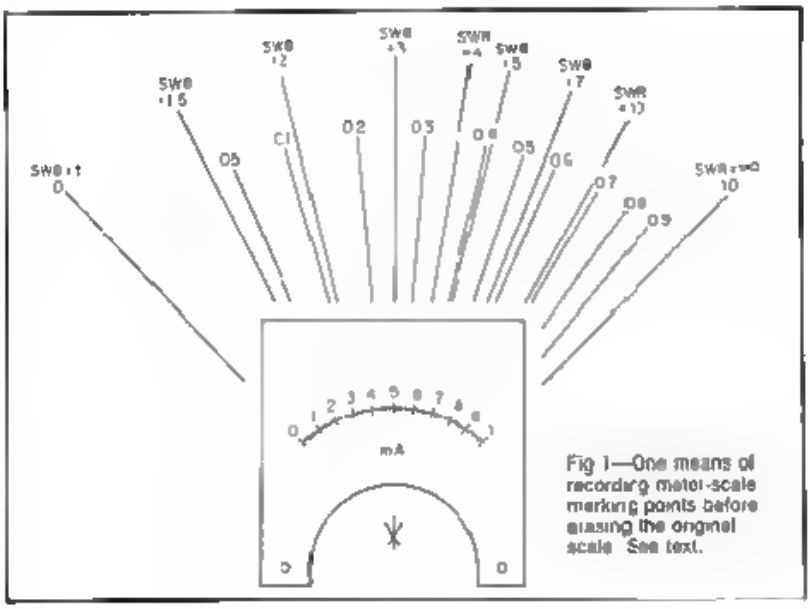


Fig 1—One means of recording meter-scale marking points before erasing the original scale. See text.

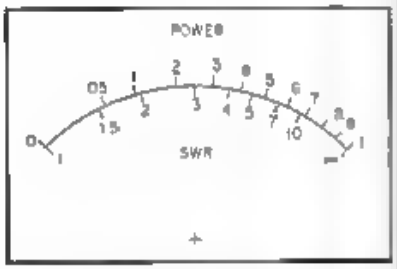
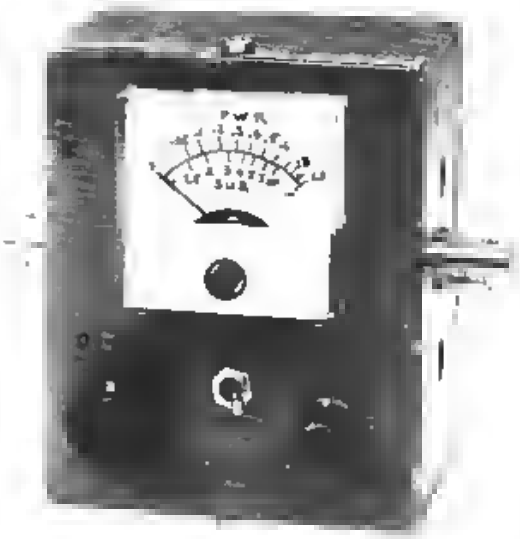


Fig 2—Example of a finished meter face. This scale is that of the meter used in "A Simple and Accurate QRP Directional Wattmeter."

From February 1990 QST, p 19:

A Simple and Accurate QRP Directional Wattmeter



Make a few small enhancements to the Bruene wattmeter and diode detector, and you have a directional wattmeter that's simple, portable, and accurate from 10 watts down to 5 milliwatts!

By Roy Lawallen, W7EL
8470 SW 152 Ave
Beaverton, OR 97007

A directional wattmeter is a really indispensable tool. Besides using it with a meter to measure SWR, you can use one to tune a home-built rig, adjust a Transmatch, measure cable loss, and a host of other things. Because it's portable, a wattmeter is an important tool in the field. With it, you can make sure the rig still works, and spot any problems with the antenna system. If you're operating QRP in Field Day or some other event, a good wattmeter can help you keep your output at five watts as the battery voltage drops.

This wattmeter, designed primarily for portable use, gives accurate readings at power levels from 5 mW to 10 W. Achieving good low-power accuracy is a bit tricky. I developed a simple correction circuit to handle the job. During the editing of this article, I learned that the technique I developed for compensating the diodes in this wattmeter's detector circuit was first discussed by John Grebenkemper, KI6WX, in his January 1987 *QST* article.[1] I encourage reading (or rereading) this excellent article.

If carefully constructed, this wattmeter should function well from below 1 MHz at least into the mid-VHF range. One prototype tested in the ARRL Lab maintains better than ± 7% of full scale accuracy, on all ranges, up to 432 MHz.

Circuit Description

A basic directional wattmeter has three major parts: directional coupler, detector, and meter circuits. Each block can be optimized for a particular application. Here's a description of each block.

[1]Notes appear at end of article.

Directional Coupler

Two types of directional couplers are commonly used by amateurs. The venerable Monimatch circuit is simple and useful for SWR measurement, but not readily adaptable as a wattmeter except over a narrow frequency range, because its sensitivity changes with frequency.[2] The Bruene circuit doesn't have this limitation, so is more suitable for our use. It's generally implemented with capacitive dividers for sensing voltage, but I chose to use transformers for this function.[3] This results in a simpler circuit that's adjustment-free. Sensitivity can be traded for insertion loss; the values chosen for this meter result in insignificant insertion loss.

Maintaining a near-50-Ω impedance on the line through the wattmeter eliminates several frequency-dependent effects. A microstripline structure is effective for this application, and is extremely simple to build, so I used that technique in this wattmeter.

Detector

Seemingly, the detector should be the easiest part of the wattmeter to design. Well, it happened again. The simplest part turned out to be the hardest. What's so hard about using a diode detector? If you don't want to know, skip ahead to the Construction section. For the truly adventurous (mathematically, that is), I've included the sidebar, "Ac v Dc: Why the Difference?"

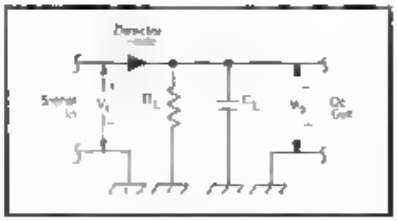


Fig 1—Simple diode detector.

Plain diode detectors, like the one shown in Fig 1, are simple and easy to use—*provided you don't require accurate results at low signal levels.* That's unfortunate, because good low-power accuracy is exactly what this wattmeter is intended to provide. Five milliwatts provides only 145 mV (peak) at the detector, so detector accuracy must be maintained down to this level. Some diodes, such as back diodes and zero-bias Schottky types, are specially designed for detecting very small signals. These, however, aren't as readily available as common silicon, germanium, and medium-barrier silicon Schottky diodes, so I investigated only the latter three types. Naturally, each has its deficiencies.

Common small-signal silicon diodes (eg, 1N914) drop too much forward voltage to be accurate at small signal levels when used with reasonable load-resistance values (up to 100 MΩ or so). Ordinary small-signal Schottky diodes are better, but still have an objectionable drop for use at low signal levels. The good old point-contact germanium diode (1N34 type) is the clear winner in this category. Applying 50 mV (dc) to a germanium diode detector produces about 45 mV at its output with a 1-MΩ load resistance. Increasing the load resistance to 10 MΩ brings the output to within 1 mV of the applied voltage.

So what's the problem? The problem is that *the results are different when you apply an ac signal to the detector!* This difference is clearly shown in Fig 2, which gives the measured output of a germanium diode detector (like the one shown in Fig 1) with three different input signals of the same peak value. On the log-log scales, the vertical spacing be-

Ac v Dc: Why the Difference?

Why does a diode detector produce less output when detecting an ac signal than a dc signal if both signals have the same peak value? Why does a pulsed-dc signal produce a lower output than a steady one? Fortunately, we don't have to look any further than the ideal-diode equation to get the answers to these questions. This equation describes the characteristics of an ideal diode—a diode that is ideal in the sense that it can be described by some fundamental principles, not in that its conduction is perfect in one direction and zero in the other.

The ideal-diode equation is:

$$I_d = I_s\left(e^{\frac{V_d q}{kT}} - 1\right) \qquad \text{(Eq 1)}$$

where

I_d = Diode forward current
I_s = Saturation current, about 10^{-14} A for silicon, 2×10^{-7} A for germanium at room temperature
V_d = Diode forward voltage
q = Electron charge, 1.60×10^{-19} coulomb
k = Boltzmann's constant, 1.38×10^{-23} J/K
T = Temperature, K

At room temperature, $kT \div q$ is about 25 mV. Note that I_s is strongly related to temperature, doubling with approximately every 10-°C rise in temperature.

Because our discussion primarily concerns small signals, let's see how the ideal diode behaves with small voltages or currents applied. The small-signal I/V characteristics of an ideal germanium diode are shown in Fig A. This is simply a graphical representation of Eq 1 over a limited range. (The graph is also valid for silicon diodes if the current scale is reduced by a factor of about 20 million.) Note that the I/V curve doesn't bend at the origin; it's a straight line. This gives us our first clue about small-signal diode operation: a straight line on an I/V graph represents a constant resistance, so at very small signal levels the diode looks like a resistor, and hardly rectifies at all. (By very small signal levels I mean somewhat less than $kT \div q$ [25 mV].) The resistance of the germanium diode is about 125 kΩ in this range (an ideal silicon diode is about 2.5 TΩ (2.5×10^{12} Ω)). At higher forward voltages, the current rapidly rises; at greater reverse voltages, the current increases, then levels out at a value of $-I_s$.

If we apply dc to the circuit of Fig 1, current will flow heavily at first, then taper off as C_L charges. Eventually, the current will simply be $V_e \div R_L$. Substituting $(V_r - V_d)$ for V_d in Eq 1 and rearranging produces an equation relating V_e and V_r:

$$V_e = V_r - \frac{kT}{q}\ln\left(\frac{V_e}{R_L I_s} + 1\right) \qquad \text{(Eq 2)}$$

or V_e and V_d:

$$V_d = \frac{kT}{q}\ln\left(\frac{V_e}{R_L I_s} + 1\right) \qquad \text{(Eq 3)}$$

To get a feel for the voltage drop to expect, look at Eq 3 with V_e = 100 mV, R_L = 1 MΩ, and $I_s = 2 \times 10^{-7}$ A. This results in a diode drop (V_d) of 101 mV. In contrast a silicon diode ($I_s = 10^{-14}$ A) would drop 403 mV (503 mV in for 100 mV out) under the same conditions, but it would have 10.1-mV drop if R_L was made 20 million times larger. As the signal level increases, the drop increases—but not in proportion, so detector accuracy improves. Increasing the load resistance helps also; at 100 mV out and R_L = 10 MΩ, V_d = 1.2 mV.

Now let's see what happens with an ac or pulsed-dc signal. Looking at Figs 2 and B, we can see that when V_i is greater than V_e, the drop is the same as if the input signal were dc. However, for part of the cycle, V_i is less than V_e. During this time the diode is reverse biased and substantial current flows to the left in Fig 1. Generally, C_L is made large enough to make the ripple on V_e very small. We can then consider V_e to be a constant value after an initial charge period of many cycles of the input signal. For any signal,

$$V_e = I_{d(avg)} R_L \qquad \text{(Eq 4)}$$

where $I_{d(avg)}$ is the average current flowing through the diode over a cycle of the input signal.

The wattmeter's detector signal is a bipolar sine wave.

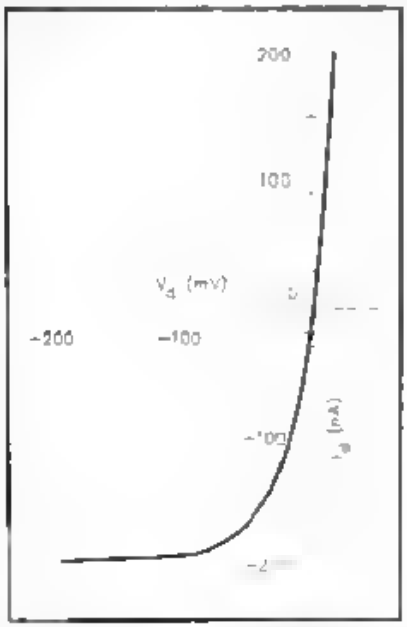


Fig A—Current-voltage characteristics of an ideal germanium diode.

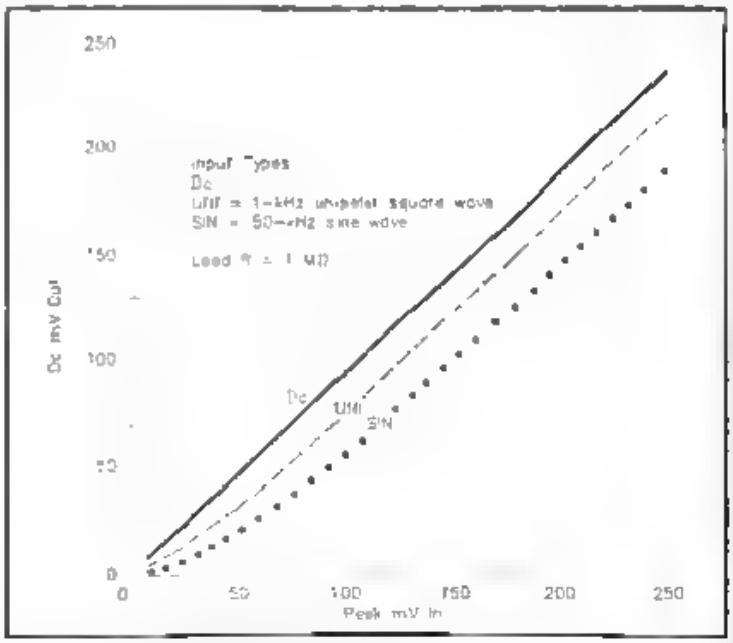


Fig B—The same data as that of Fig 2, plotted on linear axes.

but analysis of a unipolar square wave illustrates the principle and is much easier to attack mathematically (I'll discuss the sine-wave case shortly).

Consider a unipolar square wave with a positive value V_a for 50% of the cycle and 0 V for the other 50%. When the input signal is positive,

$$I_{d(h)} = I_s\left(e^{\frac{V_{d(h)}\frac{q}{kT}}} - 1\right) \qquad \text{(Eq 5)}$$

where

$I_{d(h)}$ = diode current when the input signal is high.

$V_{d(h)} = V_a - V_o$ = forward voltage (diode drop) when the input signal is high.

This follows directly from Eq 1. When the input signal is zero, the diode is reverse biased. Again from Eq 1

$$I_{d(l)} = I_s\left(e^{\frac{V_{d(l)}\frac{q}{kT}}} - 1\right) \qquad \text{(Eq 6)}$$

where

$I_{d(l)}$ = diode current when the input signal is low. $I_{d(l)}$ is negative.

$V_{d(l)} = -V_o$ = diode forward voltage when the input signal is low. $V_{d(l)}$ also is negative.

Because $I_{d(h)}$ and $I_{d(l)}$ each flow during s/2 of the input cycle, average current is found by

$$I_{d(avg)} = 1/2\,(I_{d(h)} + I_{d(l)}) = \frac{V_o}{R_L} \qquad \text{(Eq 7)}$$

Combining Eqs 5, 6, and 7 and solving for $V_{d(h)}$:

$$V_{d(h)} = \frac{kT}{q}\ln\left[2\left(\frac{V_o}{I_s R_L} + 1\right) - e^{-V_a\frac{q}{kT}}\right] \qquad \text{(Eq 8)}$$

$V_{d(h)}$ represents the difference between the peak value (V_p) of the input signal, and the output voltage (V_o). An analysis of the argument of the logarithm shows that it's always greater than the dc case. In fact, when V_o gets much greater than $kT \div q$, $V_{d(p)} = V_{d(dc)} + (kT \div q)$ in (2), or about 18 mV at room temperature. This result is independent of R_L, so even if R_L is large to minimize dc drop, the added drop due to applying ac will stay the same (as long as $V_o >> kT - q$). The added offset is clearly shown in Fig 8. Values calculated from Eqs 1 and 8 fall almost exactly on the graphs. A similar analysis for a bipolar square wave results in the same limiting value of 18 mV of excess drop.

Math is fun! But everyone's got a limit, and mine falls within the large gap between analyzing the detector with square-wave and sine-wave inputs. However, some generalizations can be made without having to do a rigorous sine-wave analysis. With a sine wave applied, one would expect diode forward current to flow for only a small part of the input cycle, resulting in a greater drop than with a square wave applied. The sine-wave case was studied with a mathematical computer model, and the results agreed very closely with the measurements presented in Figs 2 and 8. —*W7EL*

tween plots is proportional to the fractional (percentage) difference between the outputs. As you can see, a simple germanium detector is accurate for ac signals *only when the peak input signal level is above about 1 V.* A simple correction circuit is the secret to this wattmeter's accuracy. I'll explain it in a moment, but first I'll briefly explain why the detector output is less with an ac than with a dc input.

When the forward or reverse voltage across a diode gets very small (a few millivolts), the reverse and forward currents are approximately equal for a given applied voltage; that is, the diode acts like a resistor. (I measured a typical germanium diode's resistance as about 120 kΩ, a value very much smaller than that of silicon diodes.) If dc is applied to the detector, the detector circuit acts like a voltage divider, with the input voltage dividing between the diode resistance and the load resistance. When ac is applied, however, the current flow during the negative half-cycle removes a substantial part of the charge put on the load capacitor during the positive half-cycle, resulting in a lower detector-output voltage. The effect is waveshape and duty-cycle dependent, but isn't related to the frequency of the input signal.

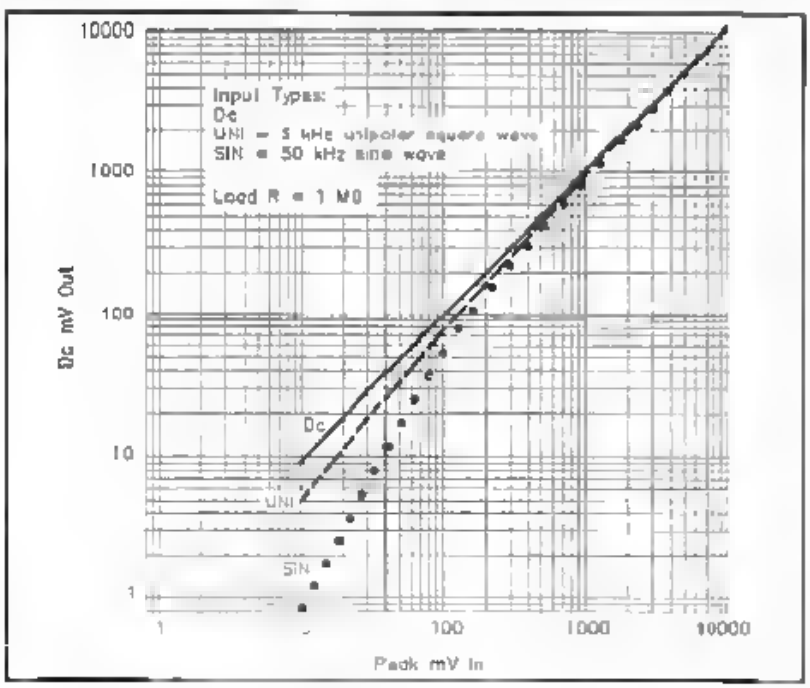


Fig 2—Dc output voltage of the detector of Fig 1 with three types of input signals.

Silicon diodes exhibit the same properties, but at different levels. Silicon-diode resistance at few-millivolt levels is about a million times larger than that of germanium diodes, but extremely large load resistances (10^{12} Ω or so) would have to be used to bring the forward drop to the germanium-diode level. The much smaller currents flowing through the much larger resistance result in the same net effect. The observed ac/dc difference is explained by the ideal-diode equation (see the sidebar). Dc and unipolar-square-wave measurements were compared to results predicted by the ideal-diode equation with extremely good agreement.

Common 1N34A germanium diodes purchased from Radio Shack® were found to be satisfactory for this detector.

The Meter Circuit

See Fig 3. An op amp is a logical choice to provide a high load impedance for the detector and a low-impedance output. Most op amps, however, have enough input-bias current to produce a significant voltage across a high-value detector-load resistor, ruining the measurement. Fortunately, operational amplifiers that have input-bias currents of only a few picoamperes—more than adequate for this application—are readily available. The CA3160 (and its externally compensated equivalent, the CA3130) has the desirable combination of extremely low input current,

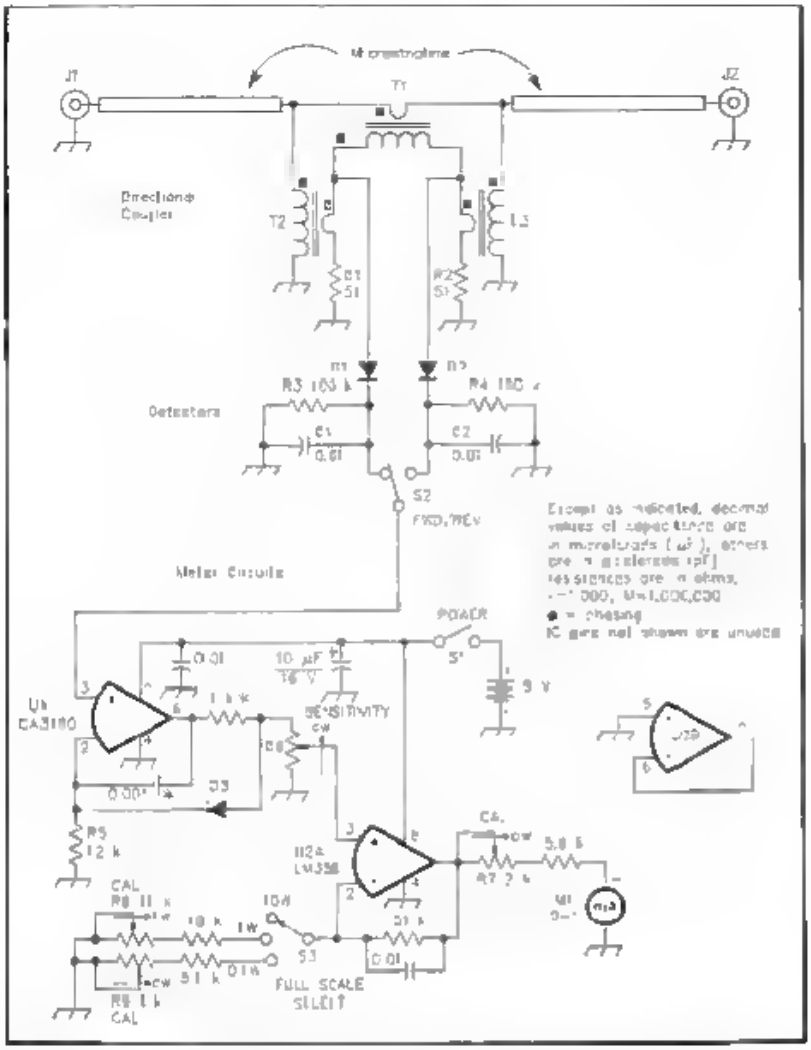


Fig 3—Schematic diagram of the QRP directional wattmeter. All fixed resistors are ¼-W, 5%-tolerance units. D1-D3 are common 1N34 germanium diodes; they should be matched (as discussed in the referent of Note 1) for best performance. T1-T3 are FT-37-72 ferrite cores with single-turn primaries (one pass through the core—see Fig 5). The primaries of T2 and T3 are comprised of the ungrounded leads of R1 and R2, respectively. Each transformer has a single secondary winding consisting of 10 evenly spaced turns of no. 28 enameled wire. Do not substitute a different core for the FT-37-72. R6 can be any value between 10 kΩ and 100 kΩ. The resistor and capacitor associated with U1) marked with asterisks may be necessary to eliminate instability in the op amp, although my prototypes don't require them. See text.

input-and output-voltage range down to the negative supply rail and moderately low current consumption. To my knowledge, no other readily available op amp shares this set of features, if you know of one, you may substitute it for the CA3160 used at U1.

The second op-amp input is used by the diode-compensation circuit, so a second stage is required to permit variable gain. The only requirements for the second op amp are moderate current consumption and the ability to handle input and output signals down to the negative supply rail (ground). The LM358 contains two such op amps, one section (U2B) is unused in this application. You could substitute one section of an LM324, another CA3160, or any other op amp having the required characteristics.

The diode-compensation circuit (D3 and R5 in Fig 3) creates an offset that approximately compensates for the drop across the detector diode. If dc was applied to detector D2/R4/C2, and if compensation resistor R5 and detector load resistor R4 were equal, perfect compensation would result (assuming that D2 and D3 were identical). However, the circuit is actually compensating the detector *ac* drop with a *dc* drop, so more current must flow through compensation diode D3. This is accomplished by making R5 smaller than R3 and R4. Although the compensation isn't perfect, it's extremely good, and a remarkable improvement for only two added components. Without the compensation circuit, the wattmeter error was 30-50% for small signals (5-50 mW); with the compensation circuit, measured error is less than 7% over the same range. In addition, the compensation circuit tracks well with temperature; an important consideration for portable use.

In John Grebenkemper's circuit (see Note 1), an additional resistor and capacitor in the feedback network are used to enhance stability in the op amp. I saw no signs of instability in my prototype, but if you experience instability problems, adding these components (marked with asterisks in Fig 3) should help.

Construction

Meter Face

You'll need to make new scales for the meter or, at the very least, add markings to the existing scale. It's fairly easy to make the new scales readable and somewhat harder to make them look nice. If you decide to make new scales rather than add marks to the existing scale, you'll want to record the correct places to make the new marks before you obliterate the old scale. See this month's Hints and Kinks column for one method of relabeling a meter face.

Directional Coupler

This wattmeter works best if there's a constant 50-Ω impedance from the input to the output, but it's not highly critical. There are always impedance bumps at the transitions

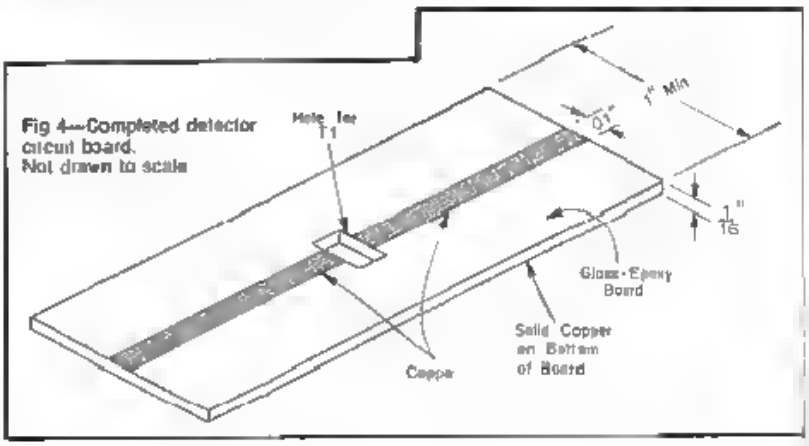


Fig 4—Completed detector circuit board. Not drawn to scale

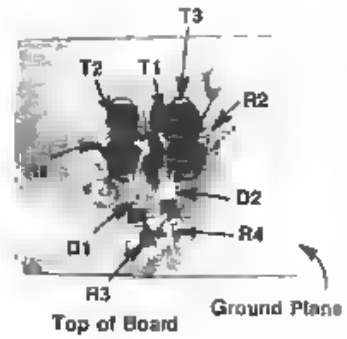


Fig 5—Close-up views of the completed directional coupler. T1 is a straight piece of insulated wire spanning the slot in the microstripline. T1's secondary winding is routed through the hole cut for T1. The top view shows the components comprising the directional coupler and detector. To achieve minimum lead lengths, R1-R4 are mounted vertically on the ground plane. Capacitors C1 and C2 are not shown, although they should be mounted directly across R3 and R4, respectively.

between the coax connectors and the microstripline, and a large bump at the coupler transformer, but these bumps can be made insignificant at HF with a little effort. Only the microstripline and directional coupler are sensitive to layout, and you have considerable latitude with these components if you know the rules.

The microstripline will be the simplest circuit board you've ever made. If you've never made a PC board before, don't worry—you can't go wrong with this one. It consists simply of a single trace on one side of the board and a ground plane on the other. There are at least three ways to fabricate this board: You can stick adhesive copper tape to the non-foil side of a single-sided board, you can etch a double-sided board, or you can cut along the edges of the trace with a knife and peel away the unwanted copper.

No matter which method you choose, start with a piece of 1/16-inch thick, glass-epoxy PC board. The board's length should equal or exceed the distance between wattmeter input and output connectors, and the board should be at least one inch wide (wider is okay). The width of the microstrip, which should be about 0.1 inch, determines the impedance of the line. Impedance doesn't change much as line width changes, and the impedance isn't too critical for this application. So if you don't have a decimal ruler, just make the trace a bit thinner than 1/8 inch—it'll be close enough. After making the board, cut a hole in the center just large enough to accommodate transformer T1. The finished microstripline should look like that shown in Fig 4.

Mount the coupler components using *short* leads. A suggested layout is shown in Fig 5. Then assemble the rest of the wattmeter and mount the completed coupler between the input and output connectors. The connections from the connectors to the microstripline must be very short, particularly the ground connections. If possible, put a solder lug or lugs on the connector-mounting screws and solder the lugs directly to the bottom of the line as shown in Fig 6.

A template package containing a PC-board pattern integrating the wattmeter circuit and directional coupler, parts-placement diagram and other information is available from the ARRL Technical Department Secretary for a no. 10 SAE with return postage for 1 ounce.

Tips

A small center-off toggle switch, wired for **REV-OFF-FWD** operation, is a convenient way to combine S1 and S2. However, small toggle switches are amazing in their ability to turn themselves on at the slightest provocation—like being bumped around in a suitcase or backpack. So if you're going to use this wattmeter for portable operation, use some other kind of switch (a slide switch, for example) for, or in series with, S1.

I've seen quite a few articles implementing the Bruene directional-coupler circuit with a powdered-iron core (eg, T-68-2). The low winding impedance of such a transformer will ruin the accuracy of this circuit, so don't use powdered-iron cores for the transformers in this wattmeter.

Adjustments

All you'll need for adjustment is a high-impedance dc voltmeter. Connect the voltmeter between the wiper of the **SENSITIVITY** control (R6) and ground. Connect a temporary jumper between pins 7 and 3 of U1. Turn R6 fully counterclockwise. Set the **FULL SCALE SELECT** switch S3 to the 10-W position. Turn on **POWER** switch S1. Slowly turn R6 clockwise. As you do, the wattmeter and voltmeter readings should increase. If not, turn the wattmeter off and check your wiring.

Adjust R6 for a voltmeter reading of 6.49 V, then adjust R7 so the wattmeter reads full scale. Adjust R6 for a voltmeter reading of 2.05 V, then switch S3 to the 1-W position. Adjust R8 until the wattmeter indicates full scale. Adjust R6 for a voltmeter reading of 0.649 V, then switch S3 to the 0.1-W position. Adjust R9 for a full scale reading on the wattmeter. Turn the wattmeter off and *remove the temporary jumper between pins 3 and 7 of U1*. This completes the calibration.

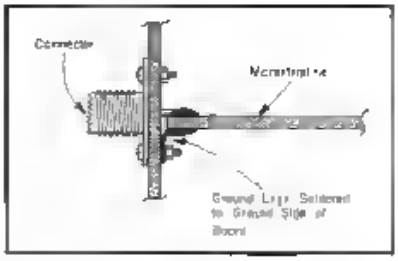


Fig 6—Using solder lugs to make a good connection to the ground side of the microstripline.

To obtain maximum reliability, measure R7, R8, and R9 and replace them with fixed resistors of the measured values; readjustment should never be necessary.

If you need to measure SWR at levels very close to 1:1, you may want to tweak the wattmeter to show a zero-reflected-power indication when connected to a resistance dummy load. The resistors to adjust are R1 and R2. Theoretically, the correct value for these resistors is 49.5 Ω each. It's not necessary to readjust the wattmeter if you change the values of R1 and R2 slightly.

Use

To measure power, select the appropriate scale and turn S6 fully clockwise. The power flowing in the line is the forward reading minus the reverse reading. To measure SWR, switch S1 to the next more sensitive setting and switch S2 to **FWD**. Adjust R6 for a full-scale meter reading. Flip S2 to **REV** and read the SWR scale. To adjust a Transmatch, put the wattmeter between the transmitter and Transmatch and adjust the Transmatch for zero reflected power.

The directional wattmeter can do anything an SWR meter can do, and many things besides. Because you can measure power anywhere in a system, you can use the wattmeter to find cable and Transmatch losses, measure transmitter power, and lots of other things. You'll be surprised how often you reach for it!

Acknowledgements

Thanks to Dave Deford, K0ED, for helping me reduce the mystery of as versus ds response of a diode detector to the realm of physics, where it belongs.

While editing this article, *QST* Assistant Technical Editor Rus Healy, NJ2L, recognized the detector-diode-compensation method as the one presented by John Grebenkemper (see Note 1). John Grebenkemper subsequently reviewed the article and made several useful suggestions, many of which have been incorporated into this article. One important consequence of John's comments is that they motivated me to model the performance of the detector/compensation combination for sine-wave inputs, during which I discovered an undesirably high sensitivity to diode saturation current, which is closely related to temperature. This resulted in lowering the values of R3, R4 and R5 to

those shown in Fig 3 from the original design values, which were much higher. John's diode measurements showed a greater variation in I_{sat} than I had found, indicating the desirability of matching the diodes as described in his article. John also pointed out the possibility of op-amp offset causing the meter to read slightly upscale with no applied signal. The worst-case error is about 2% of full-scale deflection; less in most cases. If necessary, you can add an offset null consisting of a 100-kΩ potentiometer between pins 1 and 5 of U1, with its wiper grounded.

Notes

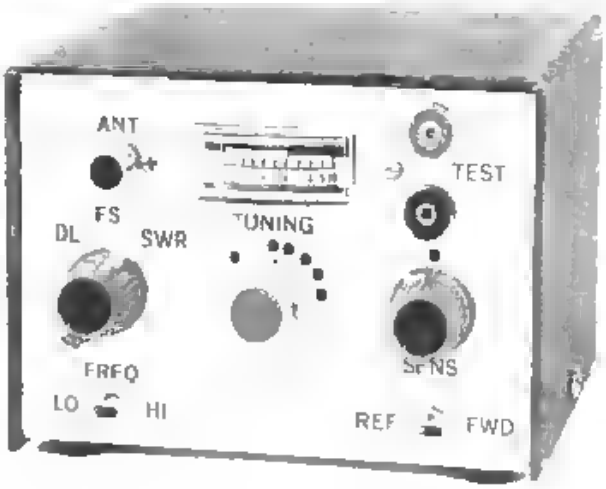
[1]J. Grabenkemper, "The Tandem Match—An Accurate Directional Wattmeter," *QST*, Jan 1987, pp 18-28. Also see "Tandem Match Corrections," Technical Correspondence, Jan 1988, p 49.

[2]This circuit was originally discussed in L. McCoy, "The Monimatch," *QST*, Oct 1958, and has been covered in many articles since.

[3]W. S. Bruene, "An Inside Picture of Directional Wattmeters," *QST*, Apr 1959.

From November 1987 QST, p 10:

Build This QRP Omni Box



Man does not live by rig alone! Combine your QRP accessories into one package for field or home use.

By Doug DeMaw, W1FB
PO Box 250
Luther, MI 49656

Do you need to carry a number of small QRP-support gadgets with you during portable operation? If so, you may be interested in how I solved my "bag-of-accessories" problem by building the most needed support units into one cabinet. A secondary advantage of utilizing these circuits is that only one panel meter and one cabinet are required. This represents a saving in dollars—an appealing fringe benefit.

You need not incorporate all of the circuits I chose for my Omni Box. On the other hand, you may prefer to add some accessory circuit that I don't find necessary for my QRP operations. The road to innovation is open to you! Whatever your pleasure, I'm sure you will be impressed with the convenience of having all of the necessary accessory items gathered together in a single housing. This is particularly handy for camping, Field Day, vacations and casual travel. Moreover, the Omni Box can be a convenient gadget for home-station use as well.

Fig 1 shows all of the circuits in my Omni Box. The instrument contains a field-strength meter, dummy load, SWR bridge, frequency standard and continuity tester.

Field-Strength Meter Section

An indication of relative field strength is helpful when checking antenna performance and patterns. This instrument may be used as a tune-up indicator, or as a relative output-power monitor. Still another application is that of a frequency meter to ensure that the transmitter is providing output in the correct amateur band. The circuit may be used also as a RF "sniffer" when troubleshooting a transmitter.

Refer to the field-strength meter circuit in Fig 1. Two operating ranges are provided. When S3 (FREQ) is open (LO), the tuning range of C1 provides coverage of 2.6 to 10.5 MHz, thereby permitting tests on 80, 75, 40 and 30 meters. When L1 is placed in parallel with T3 (HI), the effective circuit inductance is 1.5 µH. This provides coverage from 6.9 to 25.4 MHz for use on 40, 30, 20, 15 and 12 m. See Table 1.

Table 1

Approximate TUNING Dial Settings for the Field Strength Meter

FREQ LO Range

Band	C1 Setting
80 m	12:30 (o'clock)
40 m	2:30 "
30 m	5:00 "

FREQ HI Range

40 m	10:00 "
30 m	12:30 "
20 m	2:00 "
15 m	2:30 "
12 m	3:00 "

C1 is a miniature broadcast-band radio variable capacitor.[1] You may use any capacitor that provides 365 to 400 pF of maximum capacitance. The minimum capacitance (plates unmeshed) should be 20 pF or less. You may also use the variable capacitor from a transistor AM radio by placing both sections in parallel; this provides approximately 225 pF of maximum capacitance. Using this small a capacitance value will limit the tuning range of the field-strength meter, so fixed-value capacitors must be shunted across C1 to cover the low end of each range. Also, the calibration data in Table 1 will not be applicable.

The secondary winding of T3 provides low-impedance coupling to D1 and D2. The link also prevents excessive loading of the tuned circuit, and helps ensure a workable Q on both ranges (too low a Q will restrict the sensitivity of the instrument).

D1 and D2 function as a voltage doubler. The rectified RF voltage causes current to flow through the indicating meter, M1. Therefore, the greater the field strength, the higher the meter reading. C1 is adjusted for a peak meter reading, and R6 is used as a sensitivity control to keep the meter from being driven offscale. A 24-inch whip antenna connected to J1 should suffice for most field-strength tests.

Dummy-Load Section

A dummy load is important when we need to check transmitter performance or make tuning adjustments. In the dummy-load circuit of Fig 1, I use four 200-ohm, 2 W resistors (R1-R4, incl) in parallel to provide a 50-ohm load. RF voltage across the dummy load is rectified by D3 and filtered by C4. The resulting dc voltage is applied to M1 through S1. R5 isolates the dummy load from the metering circuit and makes the meter response more linear. The meter provides a visual indication of the transmitter output energy.

The meter may be calibrated in watts by applying a known power (say, 5 W) to the load and adjusting R6 (SENS) for a full-scale reading on M1. The power is then reduced in 1-W steps, and the meter reading noted at each step. These readings are logged for future use (see Table 2). I placed

[1]Notes appear in end of article

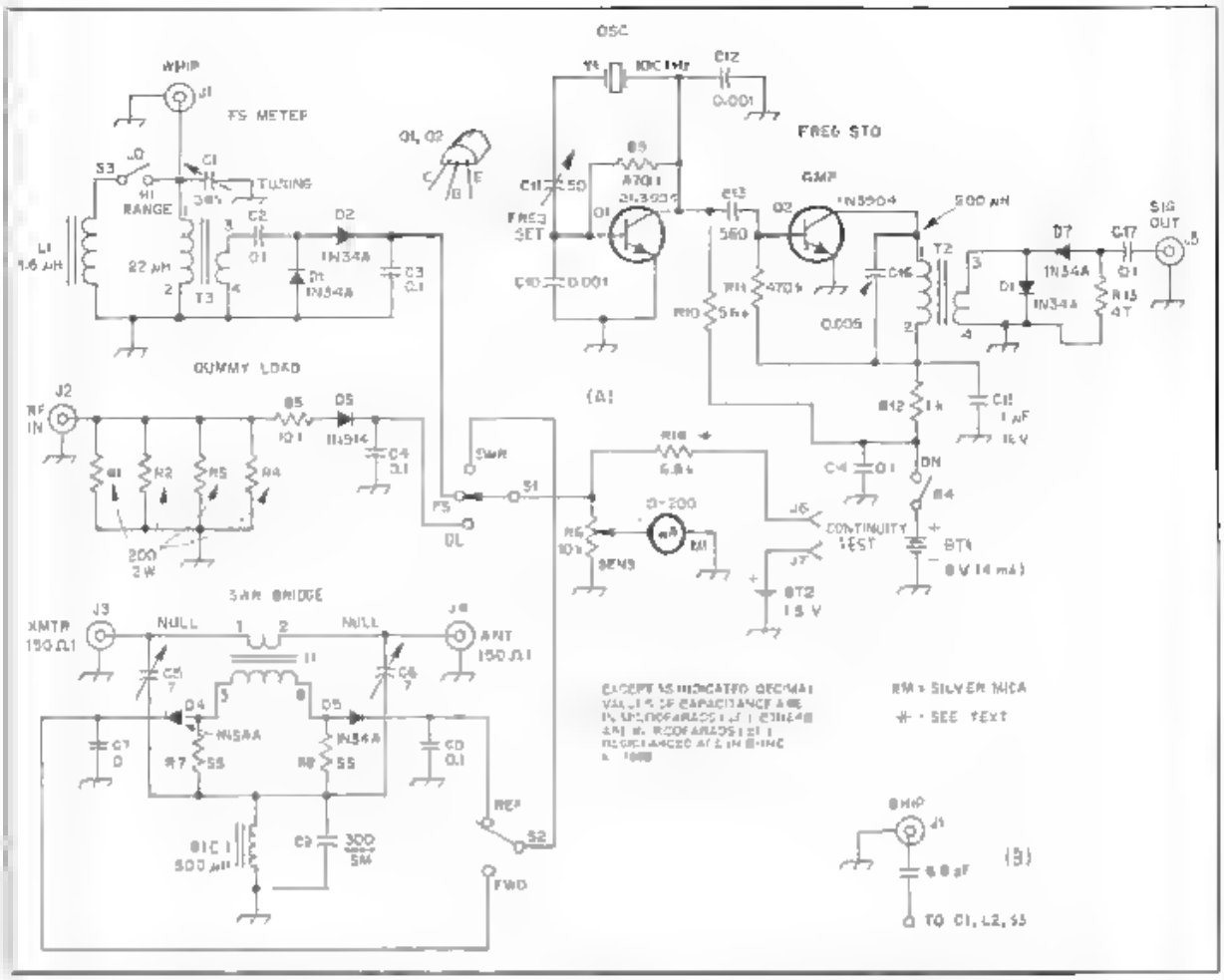


Fig 1—Schematic diagram of the Omni Box circuits. Fixed-value capacitors are miniature chip or disc ceramic types except for C15, which is electrolytic. Fixed-value resistors are ¼-W carbon composition except for R1-R4 (incl, which are 2-W units. Numbered parts that do not appear in the parts list are identified for circuit-board layout convenience.

C1—Miniature 365-pF variable (see Note 1).

C5, C6—Miniature 7-pF piston trimmer or equivalent unit with low minimum capacitance (see text).

C11—50-pF trimmer (Radio Shack 272-1340 or equiv).

J1-J5, incl—Single-hole-mount phono jack.

J6, J7—Pin jack for test leads.

L1—1.6-µH inductor, 18 turns of no. 24 enam wire on an Amidon T-50-6 (yellow) powdered-iron toroid.

M1—Miniature 200-µA dc meter (see text).

R6—Miniature 10-kΩ linear-taper carbon.

RFC1—Miniature 500-µH RF choke (1 mH also suitable).

S1—Single-section, three-position rotary switch.

S2, S3, S4—Miniature SPDT toggle or slide switch.

T1—Broadband transformer. Secondary winding is 30 turns of no. 26 enam wire on an Amidon FT-50A-61 ferrite toroid (125 µ). Primary is a two-turn winding of no. 24 enam wire over secondary winding.

T2—500-µH primary winding. Use 38 turns of no. 26 enam wire on an Amidon FT-50-43 ferrite toroid (800 µ). Secondary winding has 10 turns of no. 26 enam wire.

T3—22 µH primary winding. Use 20 turns of no. 26 enam wire on an Amidon FT-37-61 (125 µ) ferrite toroid. Secondary winding consists of 5 turns of no. 26 enam wire.

Y1—100-kHz fundamental crystal, 30-pF load capacitance, International Crystal Co. type GP. See note 3.

marks on the front panel to allow resetting of R6.

Depending on the type of SWR bridge you use in your Omni Box, the dummy load may be a part of the bridge circuit. This will simplify the project.

SWR Bridge

You have some choice in the type of SWR bridge you use.[2] You may prefer to use the resistive bridge circuit described in the referenced article. The circuit shown here is similar to the toroidal-transformer (QRO) bridge described in that article, but it is more sensitive to make it suitable for power levels from 350 mW to 25 W.

Table 2

Calibration for a 200-µA meter with SENS at Mid-scale

RF Power (W)	*Meter Scale*
5	10
4	9
3	8
2	7
1	5
0.5	3.5
0.25	1.5
0	5

D4 and D5 rectify the forward or reflected voltage (selected by S2) to provide a dc voltage for the meter. Trimmer capacitors C5 and C6 form a voltage divider with C9. These trimmers are used to null the bridge with a 50-ohm load connected to J3 or J4. A coaxial-cable jumper may be connected between J3 or J4 and J2 (dummy load) when nulling the bridge circuit.

To null the bridge, set S2 to FWD, connect the 50-ohm load to J4 and apply trans-

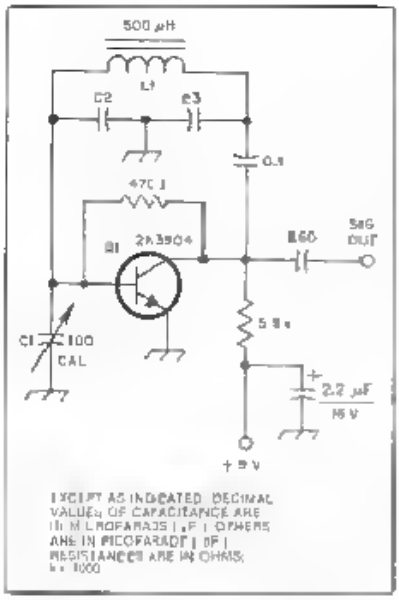


Fig 2—Suggested circuit for a 100-kHz LC oscillator. C1 is a 100-pF ceramic trimmer. L1 consists of 85 turns of no. 32 enam wire on an Amidon FT-50-61 ferrite toroid core. C2 and C3 are 0.01-µF, high-Q capacitors, such as polystyrene or Mylar® units. C1 is adjusted for zero beat the oscillator output with WWV. A coating of coil cement should be applied to L1.

result in greater sensitivity for the Omni Box functions than the specified 200-µA unit. This increased sensitivity can be particularly beneficial when using the field-strength and SWR-bridge circuits. Most imported meters with a 50- or 100-µA movement are in a conventional format, and are easier to read than the smaller, edge-reading types.

100-kHz Frequency Standard

There may be no more useful accessory than a secondary frequency standard. Many home-brew QRP transmitters—particularly those with VFOs—are prone to frequency changes as the ambient temperature varies. The problem is not limited to homemade gear. I have used several pieces of commercial QRP gear that exhibit frequency-calibration problems. Also, shock or vibration can shift a trimmer capacitor or a slug-tuned-core setting. Out-of-band or out-of license-class-segment frequency excursions can be avoided by making periodic transmitter dial calibration checks using a properly calibrated receiver. I like to know my operating frequency, so I always carry a secondary frequency standard with me on QRP expeditions.

The frequency-standard circuit in Fig 1 holds its calibration quite well. Q1 is a 100 kHz crystal-controlled oscillator. C16 and C12 are feedback capacitors that ensure circuit oscillation. These capacitors may need to be changed slightly from the values shown, depending on the characteristics of the crystal you use.

Q2 is a broadband amplifier that increases the 100-kHz energy sufficiently to permit D6 and D7 to generate strong harmonics of the crystal frequency. The diodes generate harmonics by distorting (clipping) the signal from Q2. This is particularly important when using the 100-kHz markers above 40 meters. Weak markers may not be discernible in QRN and QRM.

T2 is tuned broadly to resonance by C16. R13 provides a dc return for D6 and D7 and establishes a load for Q2. A 9-V battery supplies operating voltage for the frequency standard. It's easy to forget to turn S4 to OFF when you are not using the standard—I've done it too many times! If the switch is left in the ON position for long periods, BT1 will be depleted. It's wise to carry a spare 9-V battery with you on field trips.

Using a new 100-kHz crystal at Y1 may be costly! I suggest that you scan the surplus equipment catalogs for moderately priced 100-kHz crystals.[1] Alternatively you may use a 500- or 1000-kHz crysta

meter power to J3. Adjust R6 (SENS) for a full-scale M1 reading. Now, set S2 to REF and adjust C6 for a zero reading on M1. Next, reverse the connections—connect the transmitter to J4 and the dummy load to J3, and set S2 to FWD. While applying transmitter power, adjust C5 for a zero reading on M1. Repeat this process once more to compensate for any interaction of the two trimming capacitors.

The values of R7 and R8 are different than those in the QRO bridge in the referenced article. In addition, T1 has a two-turn link rather than having the antenna line pass through the toroidal core (the equivalent of a one-turn winding). These changes ensure greater SWR meter sensitivity, necessary for QRP use. The circuit shown may be used at power levels up to 25 W without damage to the diodes.

The Meter

A 200-µA instrument is specified for M1. There are a number of low-cost, edge-reading meters of this type available in the surplus market. Most of these are FM tuning meters, but some are calibrated for use in CB transceivers. These meters are easy to take apart for substitution of a new meter scale. A 0-10 scale that will fit most of these meters was published in the article referenced in note 2. A photocopy of this meter scale can be affixed to the faceplate of your surplus meter with rubber cement.

Using a 50- or 100-µA meter at M1 will

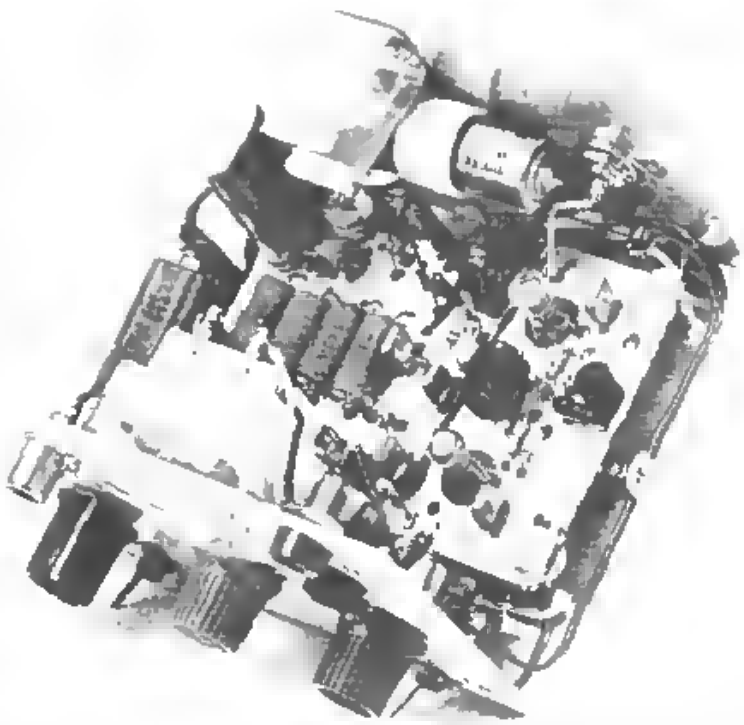


Fig 3—Interior view of the assembled Omni Box. The battery holders are affixed to the rear wall of the cabinet. The SWR bridge is at the far right of the PC board. The field-strength meter and dummy load are near the front panel at the left side of the cabinet. Y1 and the 100-kHz oscillator are located below the buttons.

with frequency dividers to obtain markers at, say, 25, 50 and 100 kHz. This approach complicates the circuit, however, and increases the current drain on BT1. Fig 2 shows an LC 100-kHz oscillator that may be substituted for Q1 of Fig 1. It will need calibration against WWV more frequently than is necessary with a crystal oscillator. It does, however, present a way to save money.

The frequency standard may be calibrated by connecting a coaxial cable between J5 and a receiver capable of receiving WWV. Tune in WWV and adjust C11 to obtain a zero beat between the output frequency of the standard and WWV. Calibration should be checked at least once a month to ensure that the standard is accurate.

Calibrate your receiver by connecting a coaxial-cable jumper between J5 and the antenna jack of your receiver. If the 100 kHz signal is too strong, you can lower the signal level by substituting a capacitor of lower value for C16 (5 to 27 pF). Tune the receiver to a convenient frequency that is an exact multiple of 100-kHz, and adjust the receiver-tuning trimmer capacitor for zero beat with the standard. Once your receiver is properly calibrated, it may be used to check the calibration of the transmitter frequency dial. A low level signal from the transmitter, such as that obtained in the SPOT position, is sufficient for calibration, and this signal level can usually be heard without an antenna.

Continuity Tester

Continuity tests are frequently necessary when we are away from our home stations with QRP gear. Situations arise where we need to check a coaxial cable or an antenna for opens or shorts. A simple continuity tester will suffice, and it eliminates the need to carry a VOM.

I added R14 (Fig 1) and two pin jacks (J6 and J7) to the metering circuit of the Omni Box. These components, along with BT2, provide a full-scale reading at M1 when a short is placed across J6 and J7. Resistances of more than 1 ohm can be observed with this tester. No switch is needed for connecting BT2 into the circuit because the line is open until the test probes are placed across a conducting path. R14 is chosen for use with a 200-μA meter. You may need to experiment with the value of R14 if you use a meter with other than a 200-μA movement. S1 may be in any position of its three positions while making continuity tests. The diodes connected to S1 block the flow of de from BT2 because their cathodes are connected toward the *positive* voltage source.

Construction Notes

Packaging of your Omni Box is a matter of personal choice. I used a Ten-Tec TG-TW 34 utility cabinet for this project. Its dimensions are 3 × 4 1/8 × 4-1/8 inches (HWD). The front and rear panels are egg-shell white, and the cover is finished in a brown wood-grain adhesive-backed plastic. The panel labels are press-on decals that were applied after the panel holes were drilled, and before the controls were mounted. Following application of the labels, I sprayed the front and back panels with Krylon® No. 1303 clear acrylic lacquer to protect the labels and give them more contrast. This product is available in office-supply stores.

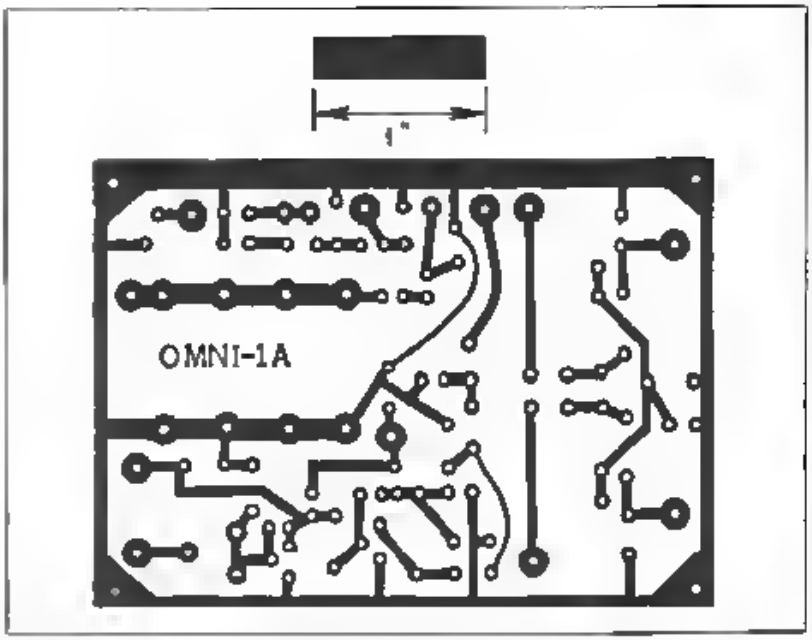


Fig 4—Circuit-board etching pattern for the Omni Box. The pattern is shown full-size from the foil side of the board. Black areas represent unetched copper foil.

An interior view of the Omni Box is shown in Fig 3. The PC board is double sided, with the copper on the component side acting as a ground plane. I suspect that single-sided board would work satisfactorily for these circuits. I used double-sided board because the input/output PC traces for the SWR bridge depend upon the ground-plane surface of the board to form 50-ohm strip lines. Elimination of the ground plane may not affect the bridge circuit significantly, because of the short distance between J3 and J4 of Fig 1. The most used controls are on the front panel of the box. S4, the ON/OFF switch for the frequency standard, is mounted on the rear panel. A U-shaped holder is used for the 9-V battery. I attached BT2 to the inner rear-panel wall with a nylon clamp. A single AA-size battery holder for BT2 would allow more convenient replacement of the 1.5-V battery. The circuit wires are soldered to the ends of BT2 in my unit.

R14 is not mounted on the circuit board. Rather, it is soldered between J6 and R6, just behind the front panel. All of the toroidal coils are mounted vertically on the PC board. I coated each of them with a homemade coil dope after they were installed. I also flowed a large drop of cement under each coil to affix them to the PC board.

I made my coil dope by dissolving small pieces of polystyrene tubing in acrylic solvent/cement. This liquid contains methylene chloride. *Warning: Do not breathe the fumes from this chemical, and avoid getting it on your skin.* A good grade of coil dope may also be made by dissolving chips of acrylic tubing or sheeting in this solvent.

A full-scale etching template for the PC board is shown in Fig 4. A parts-placement guide is shown in Fig 5. I used donut pads and PC layout tape to develop the master artwork for the PC board. I then transferred a mirror image of the pattern to a sheet of paper with a plain-paper copier. This sheet became my master artwork for Tec-200 film, from which the etch-resist pattern was ironed onto the blank PC board.[4] After drilling the holes in the board, I plated it with Kepro tin-plating solution.

Odds and Ends

The glass piston trimmers I used for C8 and C6 are set at near maximum capacitance for the desired bridge null. Had I realized this sooner, I would have substituted 6.8-pF silver mica capacitors for the trimmers. You may want to try this, assuming that the value of C9 is close to 330 pF.

Fig 1B shows a 6.8-pF capacitor in series with the line from J1. This capacitor should

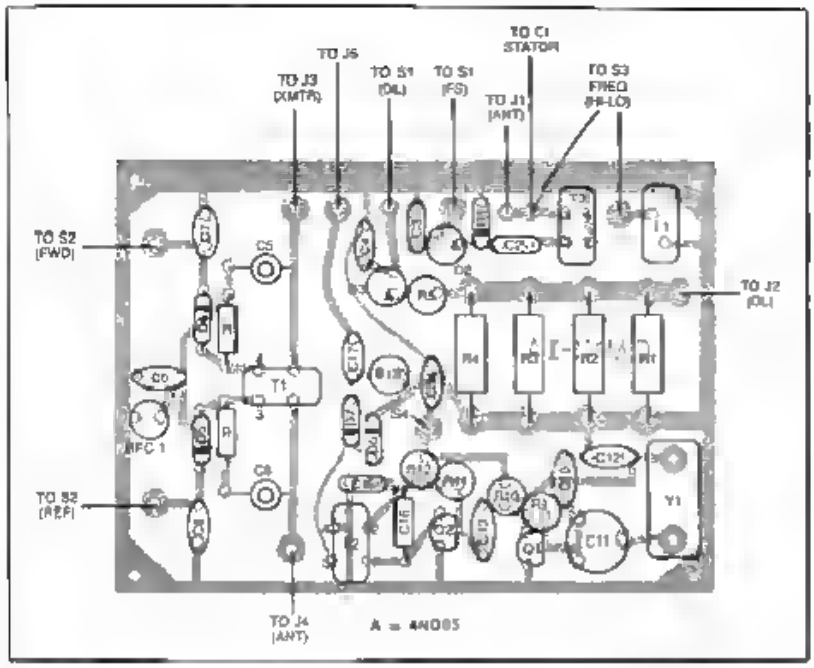


Fig 5—Parts-placement guide for the Omni Box. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern.

be added if you intend to use a longer pick-up antenna for the field-strength meter or if you connect an RF-sniffer probe to the circuit. This low-value capacitor will help to isolate the tuned circuit from the added capacitance of the probe or longer antenna. Without this change, the field-strength meter's tuned circuit will have a restricted upper-frequency range and reduced Q.

Maximum SWR bridge sensitivity (SENS set fully clockwise) is 350 mW. This is more than ample for most QRP transmitters. The dummy-load metering sensitivity may be increased by changing R5 to a lower value. The meter responds adequately at 100 mW with the value for R5 given in Fig 1.

I used an RF probe and a VTVM to measure transmitter power across a 50-ohm resistive load ($P = E_{rms}^2/R_{ohms}$). I set R6 (SENS) for a full-scale reading a MI with 5 W of RF power into the dummy load. This resulted in approximately a half-scale (12 o'clock) setting for R6. I then incrementally decreased the transmitter power and noted the readings to provide the data in Table 2. You may calibrate your meter scale for forward power by following this procedure. A scope of adequate bandwidth may be substituted for the probe and VTVM, but the resolution will not be as great as with the VTVM. You will have to convert the peak-to-peak readings of the scope to RMS values. The dummy load in the Omni Box will safely dissipate 4 W of continuous RF power. If you exceed this limit (5 to 8 W), restrict your key-down periods to 30 seconds or less, and allow a short cool-off period between tests.

You can cover the 10-m band with the field-strength meter by removing 2 turns from L1. I did not include coverage to 30 MHz because I don't operate QRP at 10 meters, likewise for 160 meters.

In the interest of miniaturization, I chose small components for most of the circuit. Surplus ceramic chip capacitors are used toward this end. Small switches are used, except for S1, which is the only suitable one I had on hand. R6 is a miniature component also.

I'm sure you will find this Omni Box as handy as I have. Maybe you'll include a QRP Transmatch in your unit to make it a complete do-everything gadget!

Notes

[1]Circuit Specialists Co, PO Box 3047, Scottsdale AZ 85257. Part No. A1-233.

[2]D. DeNaw, "The SWR Twins—QRP and QRO," *QST*, Jul 1988, p 34.

[3]JAN Crystals, 2400 Crystal Dr, PO Box 06917, Fort Myers, FL 33906-9017. Catalog no. 30, 100-kHz crystal, 0.01% tolerance, HC-13/U case, $6.50 ea.

[4]D. DeNaw, "Homemade Circuit Boards—Don't Fear Them," *QST*, Aug 1987, pp 14-16 and 22.

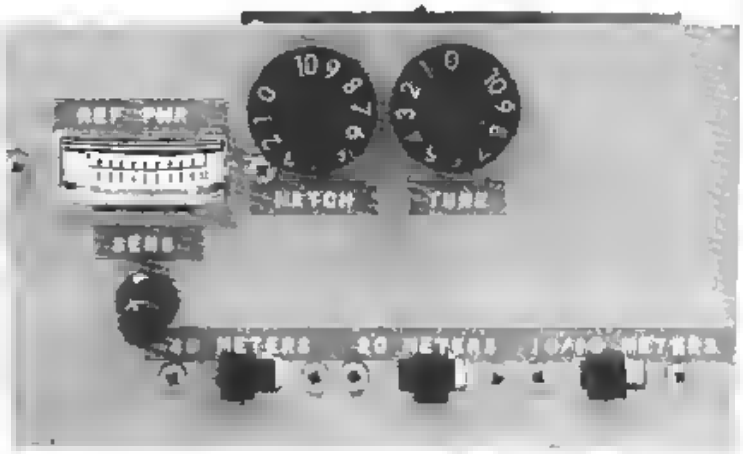
For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

From September 1998 QST, p 26:

A Simple Resonant ATU

Eliminate roller inductors and tapped coils with this simple HF-band Transmatch. This circuit is suitable for QRP or QRO.

By Doug DeMaw, W1FB
ARRL Contributing Editor
PO Box 250
Luther MI 49656



Are you weary of looking for expensive roller coils? Do tapped coils in ATUs (antenna tuning units) fail to provide the inductance resolution you need for matching a broad range of impedances? We are kindred souls if your answers to these questions are "yes." The roller-coil problem is even more acute for a QRPer: Tiny roller inductors that fit the small format of QRP gear are not available. The remaining option is a tapped coil and switch.

The circuit I shall discuss in this article is by no means new or original. The manner in which I am using it is, however, a bit uncommon. Fig 1 illustrates the circuit. Unlike other Transmatch circuits, this one is resonant at the operating frequency. Most tuners contain elements of L and C, which are used to cancel inductive or capacitive reactance in an antenna circuit. Circuit resonance is not a criterion. The popular T match that is used in most commercial Transmatches is an example of a nonresonant ATU. A resonant Transmatch offers the advantage of simplicity and harmonic reduction.

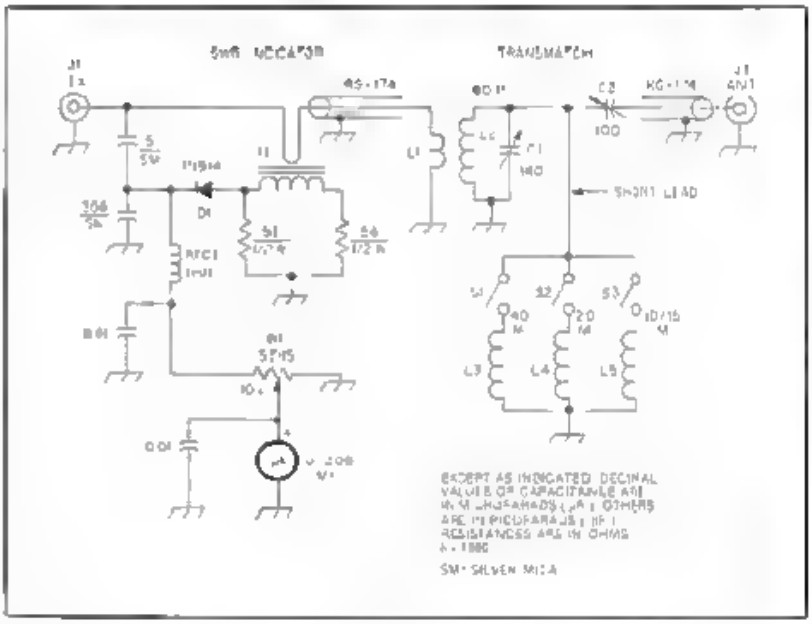


Fig 1—Schematic diagram of the SWR bridge and Transmatch. Fixed-value capacitors are disc ceramic unless noted otherwise. Fixed-value resistors are carbon composition.

C1—Miniature 100- or 140-pF air variable.
C2—10-100 pF trimmer with shaft (see note 1) or 100-pF air variable.
D1—Silicon high-speed switching diode, type 1N914 or equiv.
J1, J2—Single-hole mount phono jack or SO-239.
L1—6 turns of no. 22 insulated wire over ground end of L2.
L2—28-μH inductor. Use 70 close-wound turns of no. 22 enamel wire on a 7/8 × 2-inch length piece of PVC pipe.
L3—10-μH inductor. Use 30 turns of no. 26 enam wire, closely wound, on a 5/8 × 1-inch piece of PVC pipe.
L4—2.8-μH inductor. 16 turns of no. 20 enam wire, closely wound, on a 5/8 × 1-inch piece of PVC pipe.
L5—0.85-μH inductor. Use 9 turns of no. 20 enam wire on a 5/8 × 1-inch piece of PVC pipe. Space turns to occupy 5/8 inch.
M1—Small edgewise tuning meter, 200 μA. Surplus S meter used here with scale from page 35, July 1988 QST glued to meter face.
R1—Linear-taper, 10-kΩ potentiometer, carbon composition.
RFC1—Miniature 750-μH or 1-mH RF choke.
S1, S2, S3—SPST slide switch (see text).
T1—Toroidal transformer. Use 35 turns of no. 26 enam wire on an Amidon FT-50-61 ferrite toroid (μ_i = 125). Primary has 1 turn of no. 26 enam wire.

A Closer Look at the Circuit

Please refer to Fig 1. The main part of the circuit is L1 and L2, along with C1. Here we have a standard tuned circuit or resonator. L1 is the coupling link into the tuned circuit. As shown, C1 and L2 form a resonant 80-meter circuit. C2 has been added to permit matching the signal source (transmitter) to the load. A matched condition will prevail at some setting of C2. This is a very old trick that has been with us for decades.

There is considerable interaction between C1 and C2, since the greater the capacitance at C2, the less capacitance we need at C1 to maintain tuned-circuit

resonance. In other words, the C2 capacitance add to that of C1. For this reason we must adjust C1 and C2 alternately as we tune for minimum SWR, just as with conventional ATUs.

How do we solve the problem of multiband operation? A simple solution is provided by adding L3, L4 and L5. These coils are switched in parallel with L2 by means of S1, S2 and S3. A single-pole, three-position wafer switch can be used in place of the individual switches, although it would limit the flexibility of the circuit. I will discuss this later. As is the situation when we place resistors in parallel, coils that are placed in parallel have a net value that is less than that of the smallest coil in the combination. Therefore, we simply add L3 to the circuit for 40-meter operation, L4 for 20 meters and L5 for 10- and 15-meter operation. The 30 meter band can be covered in the 40-meter range, and 12 meters falls into the 10-15 meter range.

The advantage in placing the smaller coils in parallel with the large one is that the L1/L2 turns ratio remains the same as when only the main coil is being used. L1 can be eliminated by tapping the coil six turns above the grounded end. I chose the link method because it is easier to deal with than a coil tap. I wanted to avoid the potential of shorted turns with small wire.

The main coil has an inductance of 28 µH. The effective circuit inductance is 7.5 µH when L2 and L3 are in parallel. L2 + L4 = 2.4 µH and L2 + L3 = 0.82 µH. If all four coils are placed in parallel the net inductance becomes 0.6 µH. The singular coil inductances are given in the Fig 1 caption.

SWR Indicator

You may eliminate the SWR-sensing circuit in Fig 1 if you have a separate SWR meter to use with this tuner. I included this circuit for my convenience when operating afield with QRP equipment. I did not include the circuitry for reading the forward power. My concern is for obtaining a matched condition between the transmitter and the antenna. Therefore, I need only the reflected-power information. T1 samples the RF current (reflected). D1 rectifies the current and produces a dc voltage that is indicated at M1. The ATU is adjusted for minimum needle deflection at M1. R1 is a sensitivity control that prevents the meter from reading off scale during tuner adjustments. The SWR bridge is designed for QRP operation, as shown. A transmitter power output of 1 watt or greater will provide full-scale deflection at M1.

Construction Data

Fig 2 shows the first-cut construction detail of the coil subassembly. You will note the presence of two shaft-driven compression trimmers. I later changed C1 of Fig 1 to a small APC style air variable. This was done to eliminate mechanical problems that resulted in very "touchy" adjustment of C1. The trimmers are 10-100 pF units with 1/8-inch OD shafts.[1] I drilled holes in the ends of two ¼-inch wooden dowel rods, then glued the trimmer shafts into the dowel rods with epoxy cement. This allowed the use of standard knobs with ¼-inch holes.

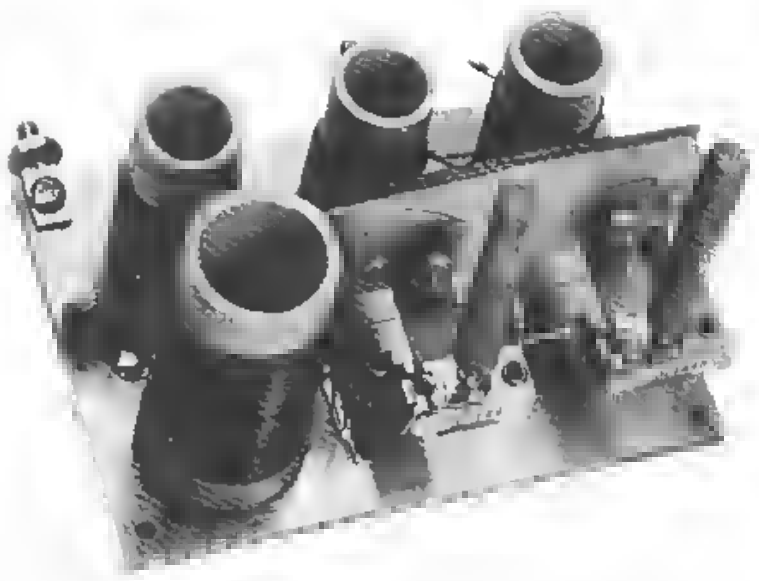


Fig 2—The ATU subassembly before C1 was replaced with a small air-variable capacitor. Holes are punched in the base plate (PC board material) to accommodate the OD of the coil forms. The coils are cemented into the holes with epoxy glue. A small shelf is soldered to the base plate to allow mounting C1 and C2. A plastic block insulates C2 from ground. Wooden dowel rod is glued to the capacitor shafts to allow the use of knobs with ¼-inch holes. The ends of the dowels are ground down to a true ¼-inch diameter.

Schedule-40 PVC tubing is used for the coil forms. PVC is not suitable for high-power use since it will heat and melt in the presence of high RF voltage. PVC is entirely acceptable for power levels under 50 watts. L2 is wound on 3/4 inch PVC pipe, which has an OD of 7/8 inch. The remaining coils are wound on 1/2-inch PVC pipe (5/8 inch OD). All of the coils are mounted on the subassembly base plate by gluing them into holes (5/8 and 7/8 inch diameter) that are cut in the PC-board base plate. Epoxy cement is good for this purpose. The coils are spaced apart 1 inch, center to center. The base plate is made from double-sided PC board (2½ × 3¼ inches). The grounded ends of the coils are soldered to the base plate.

Fig 2 shows a 1½ × 2-inch shelf upon which the trimmer capacitors are mounted by means of metal L brackets. A plastic insulator is bolted to the shelf to allow C2 of Fig 1 to be isolated from ground. The PC-board shelf is soldered to the base plate, and a small triangular PC board bracket is soldered between the bracket and base plate (at each end of the shelf) to strengthen the shelf. Two no. 6 spade bolts are used to affix the subassembly to the main chassis of the ATU. You may use brass or aluminum for the base plate and shelf if you prefer.

I made my chassis and panel from PC board material. The sections are soldered at the joints to form the main frame. The assembled unit is shown in Fig 3. The dimensions are (HWD) 3¼ × 5¼ × 3 inches. A 1 × 5-3/8 strip of PC board is soldered across the back of the chassis to contain J1 and J2 of Fig 1. Two strips (½ × 3 inches) are used at the sides of the main frame to serve as panel braces. I polished the copper on the PC-board material, then coated it with clear lacquer to prevent tarnishing. The panel is sprayed with gray automotive primer paint. I first sanded the panel to provide a rough surface. This helps the paint to adhere better than it would on the smooth surface. Gray Dymo™ tape labels identify the control functions. Four adhesive-backed rubber feet are affixed to the bottom of the chassis.

I used a technique that some call "ugly construction" when I built the SWR circuit. A neater job will result if you assemble the parts on a PC board, although the performance will be the same. I used a multilug terminal strip to contain most of the SWR bridge parts. Other components have mid-air joints.

I used inexpensive slide switches for S1, S2 and S3 of Fig 1. Miniature toggle switches may be substituted, or you may prefer to use a single rotary switch, as dis

[1]Notes appear at end of article

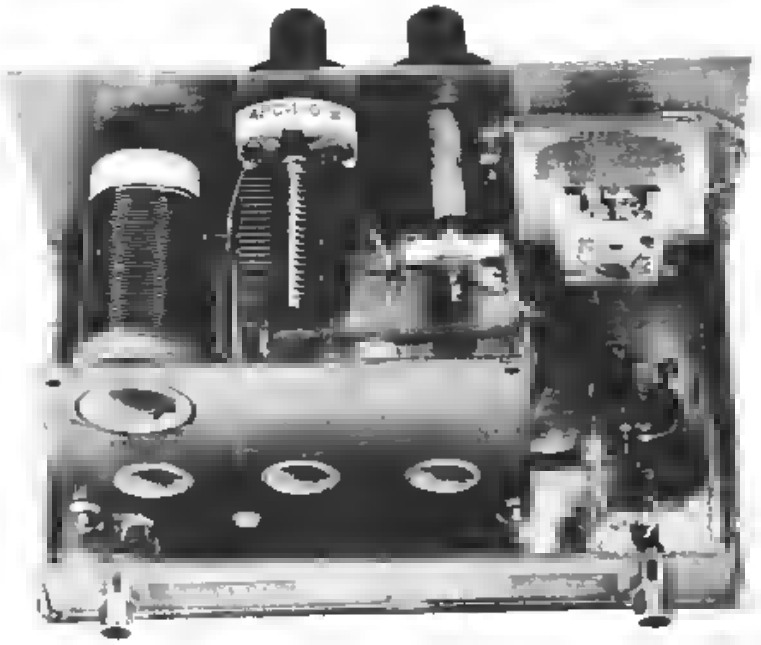


Fig 3—Interior view of the assembled ATU. "Ugly construction" (see text) is used for the SWR-bridge circuit (lower right). The subassembly, chassis and panel are made from pieces of single- and double-sided PC board. The seams are soldered to hold the sections together.

cussed earlier. Trimmer C2 may be replaced with a 100- or 140-pF air variable. If this is done, you will need to isolate the stator and rotor from ground. The circuit will function satisfactorily if you use two 100-pF capacitors (C1 and C2).

Circuit Performance

I tested this ATU at power levels from 1 to 15 watts. I used resistive loads from 15 to 1000 ohms, and obtained an SWR of 1:1 in all cases. No arcing occurred at trimmer C2. I later connected the ATU to my 80-meter dipole (coaxial cable feed) and ran it through its paces from 80 through 10 meters. Despite the complex impedance the feed line presented above 3.5 MHz, I was able to obtain an SWR of 1:1 on all bands.

Adjustment is done by setting the coil switches for the proper amateur band. With RF power applied to the circuit, adjust C1 for the lowest SWR attainable. Next, adjust C2 slightly and readjust C1 for minimum indicated SWR. Repeat this process until the SWR is 1:1. *Caution*: Use the greatest amount of capacitance possible at C2, consistent with a 1:1 SWR. Although smaller values of capacitance at C2 will result in an SWR of 1:1, the loss through the ATU increases at such settings. All Transmatches introduce some loss, but it is insignificant (less than 1 dB normally) for the most part.

Some Final Thoughts

Keep all RF leads as short as you can. This will prevent unwanted stray inductance, which can lower the tuned-circuit Q. Long RF leads such as those marked "RG-174" in Fig 1, should be made from coaxial cable. RG-174 is miniature coaxial line that is suitable for short runs and for power levels up to 40 or 50 watts at the lower amateur frequencies.

There is no reason why the circuit of Fig 1 can't be adapted for high-power use. The coils would need to be made with large-diameter wire, and the coil forms should have good high-voltage, low-loss properties. Lexan™ or fiberglass tubing[2] is good material for the coil forms. Surplus ceramic coil forms are also suitable. C1 and C2 of Fig 1 must have wide plate spacing for high power, since substantial RF voltage is present at the top of L2. S1, S2 and S3 need to be high-quality RF ceramic switches if QRO use is contemplated. Fair Radio Sales in Lima, Ohio sells surplus RF power switches.[3]

You may use toroidal coils for your QRP ATU. This will enable you to make the tuner smaller. For example, L2 would have 35 turns of no. 24 enamel wire on an Amidon FT-82-63 core. L1 would consist of 3 turns of no. 24 wire over L2. For L3 use 24 turns of no. 24 wire on an FT-50-63 toroid. L4 would have 23 turns of no. 24 wire on an Amidon T-50-2 toroid, and L5 would consist of 15 turns of no. 24 wire on a T-50-6 core. There is no reason why you can't design a PC board that can contain the four toroidal coils, plus the SWR bridge. This would result in a low-profile, compact ATU.

I wrote this article in order to share some old ideas that may have been forgotten by some of you. I hope you have found the circuit and construction hints interesting and useful.

Notes

[1]Trimmers with shafts are available from Hosfelt Electronics, 2700 Sunset Blvd, Steubenville, OH 43952. Sales line: 800-524-6464 (catalog available).

[2]Plastic rod, tubing and sheeting (many types of plastic) are available by UPS or truck line from U S Plastic Corp, 1330 Neubrecht Rd, Lima, OH 45801. Sales line: 800-537-9724 (catalog available).

[3]Fair Radio Sales, Inc, Box 1105, 701 E Eureka St, Lima, OH 45802, tel 419 227-6573.

From 1990 ARRL Handbook *Ch 34, p 19:*

A Balanced QRP Transmatch

The balanced QRP Transmatch shown in Figs 39 through 41 was designed and built by Zachary Lau, KH6CP, of the ARRL Lab. It is designed for use with balanced feed lines, although random wire antennas can be fed if one of the antenna terminals is grounded. Unlike most Transmatches designed for use with balanced feed lines, this design features a balun at the input, rather than at the output. As a result, the balun sees impedances close to the design impedances once the Transmatch has been properly adjusted. This results in lower loss and freedom from core saturation at low power levels.

Since it is balanced currents that prevent feed-line radiation, this circuit was designed to balance currents rather than voltages. Some antenna systems use circuits that provide balanced voltages, making it necessary to make the system symmetrical in an effort to balance the currents. By going straight to balanced currents, instead of balanced voltages, it is possible to use a much simpler matching network. In addition, the actual current balance in typical amateur open-wire feed lines should be improved.

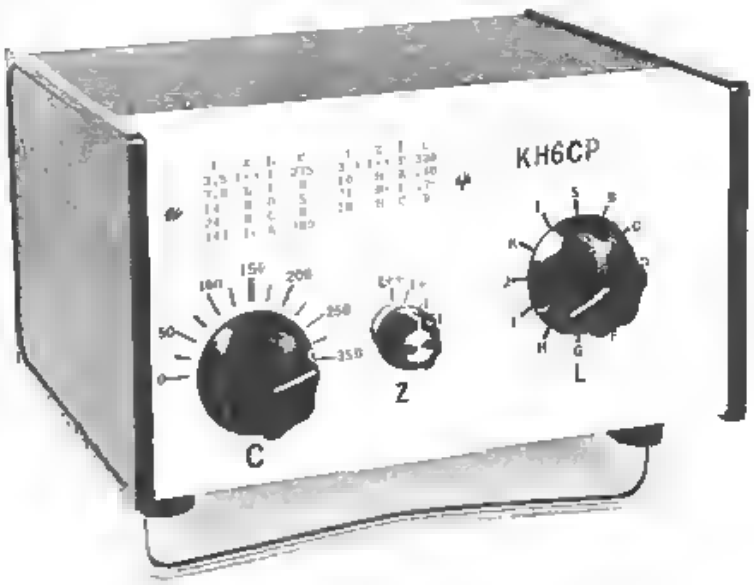


Fig 39 — The QRP Transmatch for balanced feed lines features a balun at the input of the matching network.

Construction

The inductor L1 and the capacitor C1 should be of the highest quality obtainable for best performance. Low-impedance loads will require a good inductor for efficient matching, while high-impedance loads will require a good capacitor.

L1 was wound with tinned copper wire to make it easier to adjust taps. It is necessary to wind the wire with spaces between turns to prevent shorts which may make the inductor lossy, no. 16 wire is heavy enough to stay in position on the toroid. The inductor used had a full-inductance Q of 420 at 7.9 MHz; the Q was 410 after the taps and switch were added. The use of clip lead taps is not recommended as they increase losses, although they may be useful in initially setting up tap positions.

Capacitor C1 should have a value of at least 250 pF, and larger capacitors will work even better, increasing the range of the Transmatch at low frequencies. Suitable capacitors are usually available at hamfests. The value of C2 and C3 should equal C1, and C4 should be twice the value of C1. If the calculated values of C2, C3 and C4 are not available, smaller values may be used.

Capacitor C1 must be insulated from the chassis, so it was mounted using ¼-inch Plexiglas[3] with tapped screw holes. An insulated shaft coupler was used to prevent high voltage from appearing on the knob set screw. The cabinet is a Ten-Tec MW-5 with a model 91-20e matching tilt-up bail. Although it is large for a QRP Transmatch, the cabinet matches the author's QRP rig and allows the controls to be spaced apart for easy use. The logging scale is type written paper attached to the cabinet with a Plexiglas sheet.

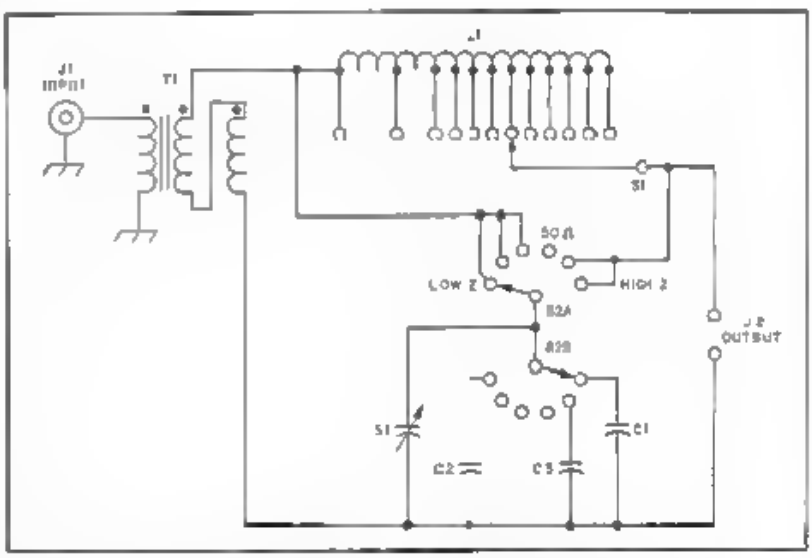


Fig 40 — Schematic diagram of the QRP Balanced Transmatch.

C1—330 pF variable capacitor, 500 V rating.
C2 C3—330 pF silver-mica capacitor, 500 V rating.
C4—660 pF silver-mica capacitor, 500 V rating.
J1—Female chassis-mount BNC connector.
J2—Two ceramic feedthrough insulators.
L1—36t no. 16 tinned wire on an Amidon T-200-6 core.
S1—Ceramic rotary switch, single wafer, 1 pole, 12 position.
S2—Ceramic rotary switch, single wafer, 2 pole, 6 position.
T1—12 bifilar turns on an Amidon FT-114-61 core, primary no. 16 enam wire, secondary no. 18 enam wire.

T1 is a bifilar-wound transformer. Winding details are shown in Fig 40. It is possible to wind this coil with only two windings, eliminating the solder joint. This coil should be duplicated exactly with regard to the number of turns and core

material unless the transformer can be tested at the operating frequency. Testing can be done by hooking up two baluns in series and measuring the insertion loss. The matching network will compensate for a poor balun, but efficiency will probably suffer. A toroidal choke balun would be recommended for a higher power version.

Switches S1 and S2 should be ceramic. Phenolic switches are not recommended, although they should work at low power levels on the order of a few watts. The switch positions should never be changed while more than a few watts of RF is applied.

Adjustment

Adjustment of the Transmatch is much easier if the approximate impedance of the load is known. In his article in *The ARRL Antenna Compendium, Volume 1*, "Mr. Smith's 'Other' Chart and Broadband Rigs" Roger Ghormley, W0KK, details how parts values for L networks can be calculated. Alternately, received signals can be peaked up by first adjusting the inductor and then the capacitor. As with any Transmatch, low power should be used in the initial adjustment. The actual power handling capability will depend on the load. The capacitor breakdown voltage is the limiting factor on high-impedance loads; a 2000-ohm load will cause the 500 V capacitors to reach their maximum rating at 62.5 W, while the maximum rating will be reached with 625 watts into a 200-ohm load. The current-handling capability of the wire is the limit on low-impedance loads; a 40-ohm load will cause a 90-W signal to generate 1.5 A through the wire, while 450 watts will generate 1.5 A if the load is 200 ohms. These values are for resistive loads; a reactive load would require higher current and voltage ratings. The unit shown here has worked well in low-power operation (up to 4.5 watts).

Fig 41 — Interior layout of the QRP Transmatch. The variable capacitor is mounted on a Plexiglas block.

From January 1985 QST, p 39:

Variable-Notch Filter for Receivers

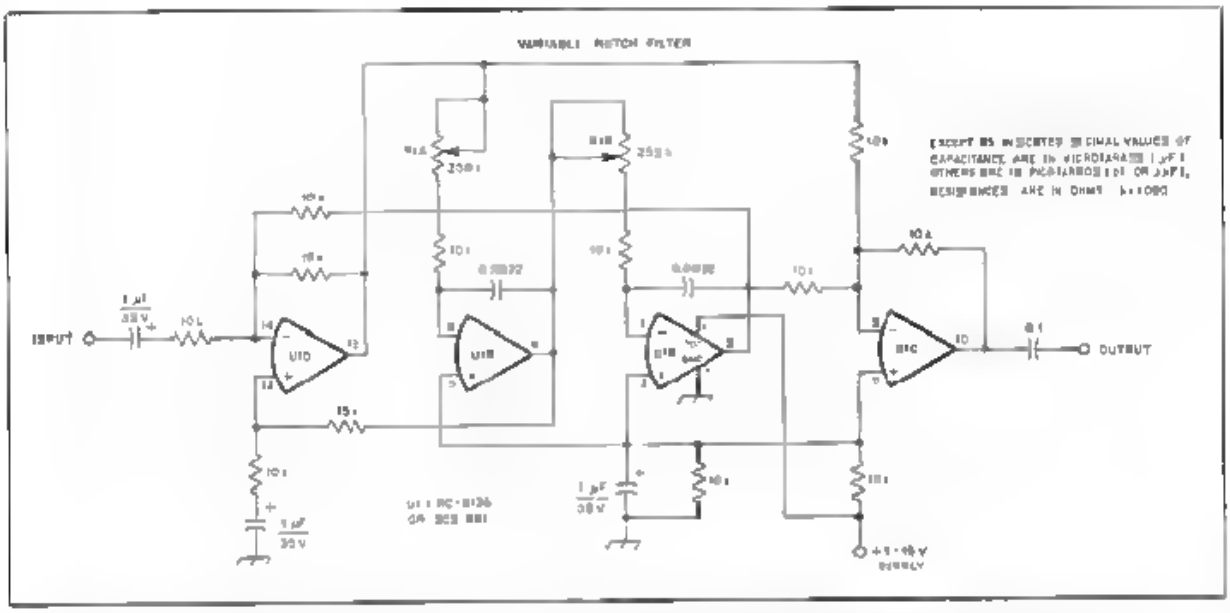


Fig 1 — Schematic diagram of the variable notch filter that K4YIZ installed in his Kenwood TS-530. The same circuit should prove useful for other receivers. R1 is a dual 250-kΩ linear taper potentiometer, such as a Clarostat DS3C1-250K-S. U1 is an RC-4136, ECG 997 or equiv quad op amp.

☐ One night before CW net time, I was fooling around the specified frequency using my Kenwood TS-530. I came across some kid giving his finals a "life test." This prompted me to think about how nice it would be to have a notch filter in the '530 similar to the one in my Ten-Tec Argosy. A variable-notch filter can be quite effective for listening to a signal that is very close to a strong, interfering station.

I checked the manual for my Argosy and found that Ten-Tec uses a simple circuit that employs a quad op amp IC and a few resistors and capacitors. A dual-section, 250-kΩ, linear-taper potentiometer serves as the tuning control. Based on my study of the Ten-Tec circuit, I decided to use it as a basis for my modification. See Fig. 1. A circuit-board etching pattern is given in Fig. 2, and Fig. 3 is a parts-placement diagram. Notice that all resistors are mounted on end to save space and to allow for a smaller board layout. I mounted my circuit inside the TS-530 and connected it as shown in Fig. 4.

Since I never used the RF gain control on my rig, I replaced it with the notch-filter potentiometer. The Clarostat DS3C-250K-S potentiometer I used is a tight fit in the chassis hole. I had to enlarge the hole slightly to fit the bushing.

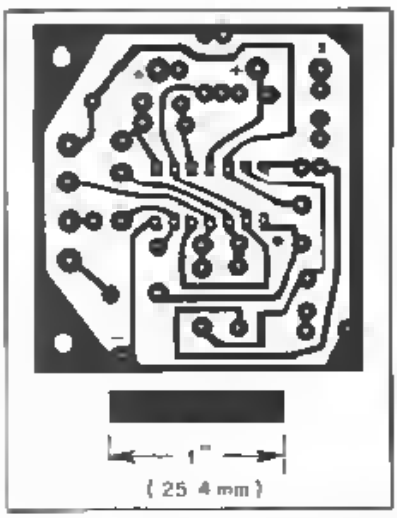


Fig 2 — Full-size circuit-board etching pattern for the notch filter, shown from the foil side. Black areas represent unetched copper.

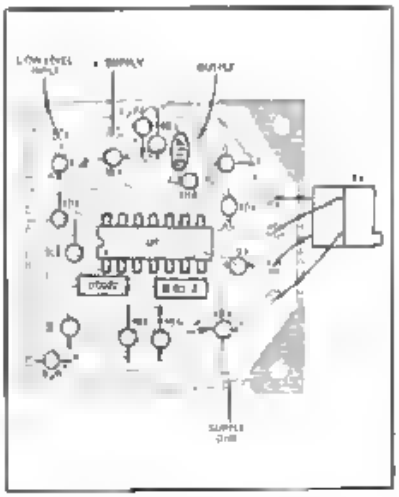


Fig 3 — A parts-placement diagram, shown from the component side of the board. Gray areas show an X-ray view of the copper pattern.

on the new control. The RF gain potentiometer is a 10-kΩ unit that I replaced with a fixed resistor.

I filed a flat on the shaft of the new control so the original Kenwood knob would fit and make the new control look like it belongs. I did not try to relabel the front panel to indicate the function of the new control. That way, I can return the rig to original form, should I ever wish to.

To mount the filter board in my Kenwood TS-530, I replaced a screw found near the edge of the audio board, between plugs 8 and 9, with a longer one. A few washers help space the filter board from the chassis. I obtained power for the filter at test point 6 on the audio board. There is a small, red coaxial cable coming from the top side of the rig and going behind the VFO. This cable carries the audio from the detector to the audio amplifier. Cut this cable and connect it to the input and output pads on the filter board, as shown in Fig. 4.

Four small wires connect the filter to the dual potentiometer on the front panel. You will have to remove the screws holding the audio board so you can lift it out of the way while replacing the RF gain control. The front panel will have to be removed to get at the nut that holds the potentiometer in place; that involves removing four screws.

With the control turned fully clockwise, the notch frequency is about 2800 Hz; it is about 300 Hz when fully counterclockwise. Both these frequencies are nearly out of the Kenwood audio system passband. When you don't need the filter, just set it to one end or the other.

I have found this to be a worthwhile project and a handy addition to my rig, especially for CW operation. You should be able to complete the modification in an evening or two. Actually, the task sounds worse than it is! It took more time to type it up than to perform the operation. One word of caution: Wire the new control so that maximum resistance on both potentiometer sections occurs when the shaft is rotated counterclockwise. — *Tom Desaulniers, K4VI2, POB 1026, Leeds, AL 35094*

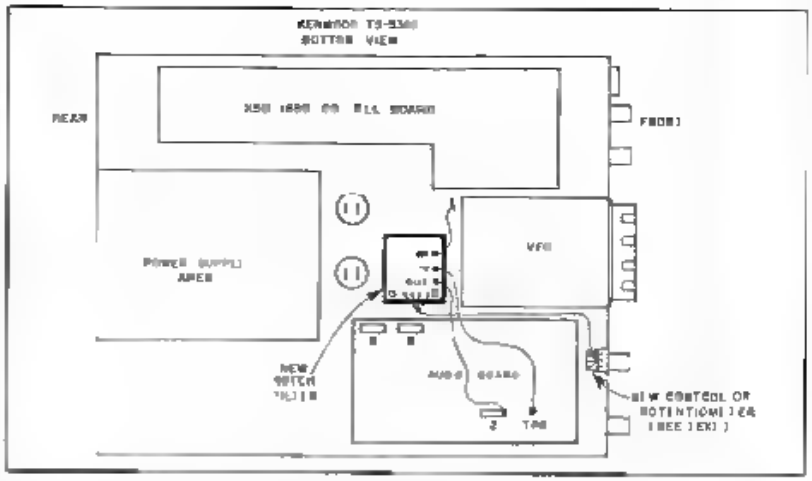


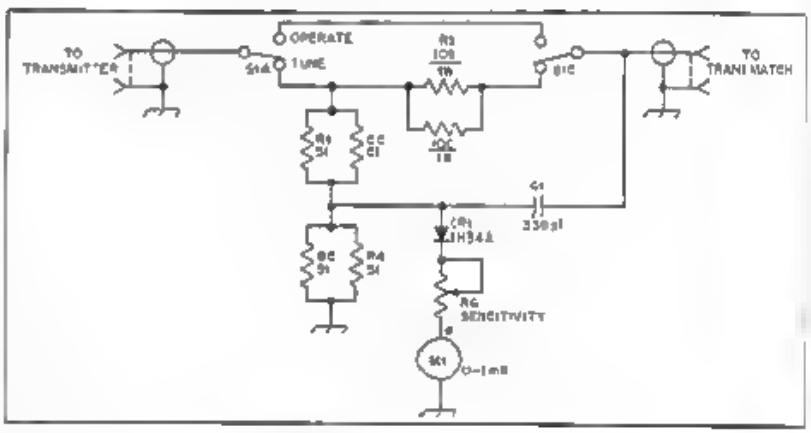
Fig. 4 — Inside view of the Kenwood TS-530 showing the notch filter wiring and mounting.

From April 1977 QST, p 57:

Simplified Output Metering Protects QRP Transmitters

After destroying a few transistors while tuning QRP transistors into a miscalculated load, I decided I needed some way to indicate proper transmitter adjustment, and then protect the rig while the antenna tuner was adjusted. An adaptation of the simple resistive SWR bridge described in the ARRL *Handbook* provides me with a dummy load, relative power-output indication and a safe method of tuning the transmitter.

As shown in the schematic diagram, the input divider (R1-R4) has a total resistance of 50 ohms. Four 1/2-watt composition resistors safely dissipate the output of my transmitter when S1 is in the TUNE position. Meter M1 indicates relative power applied to the load. The antenna is connected (through a Transmatch) and the antenna tuner is adjusted for minimum deflection on S1, or lowest SWR. R5 acts as an attenuator and effectively isolates the transmitter from the antenna, preventing possible damage to the output transistor of the rig. When the SWR has been reduced to its minimum, S1 is placed in the OPERATE position. M1 now indicates relative power output into the antenna. CR1 may be any germanium signal diode. C1 is either a ceramic-disc or silver-mica capacitor. S1 should be a ceramic rotary switch (dpdt), although a phenolic rotary switch or a slide switch is adequate for use on the 80-meter band. *Albert S. Woodhull, N1AW, ex-WIGSJ, POB 343, Amherst, MA 01004*



Protective circuit for QRP transmitters.

From June 1985 QST, p 39

An Accurate, Inexpensive Frequency Marker

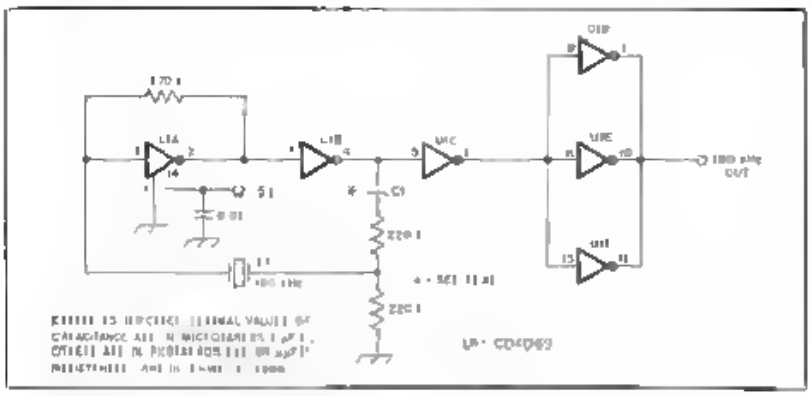


Fig 3—A schematic for the frequency marker. U1 is a CD4069, or equivalent, CMOS hex inverting buffer. Y1 is a 100-kHz CX-1H crystal from Statek.

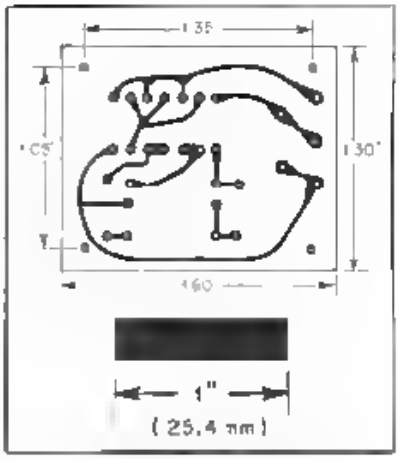


Fig 4—Full-size circuit-board etching pattern for the frequency marker, shown from the foil side. Black areas represent unetched copper.

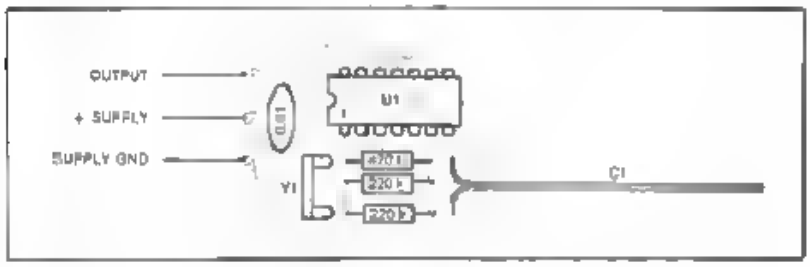


Fig 5—A parts-placement diagram shown from the component side of the board. Gray areas show an X-ray view of the copper pattern.

☐ A highly accurate series oscillator can be constructed from seven components for about $10. The marker is based on the model CX-1H quartz crystal, which is a tuning-fork resonator manufactured by the Statek Corporation. Statek produces these crystals for use in quartz watches and they have a frequency tolerance of ±0.005% at 25° Celsius. The modules cost about $6 each.

A schematic of the frequency-marker circuit is shown in Fig 3. It consists of three cascaded inverters and three parallel buffer inverters. The entire circuit can be etched on a 1.6- × 1.3-inch board, as shown in Fig 4 (Fig 5 is a parts-placement diagram for the circuit.) I used the components specified by Statek and the marker oscillates at 100.005 kHz. (Larry Wolfgang, WA3VIL, used a "gimmick" capacitor for C1 in the frequency marker he built in the ARRL lab. The gimmick is two wires of a no. 24 AWG ribbon cable. Start with the wire somewhat longer than 3½ inches, connect the output of the frequency marker to a frequency counter, and trim the gimmick wires until the marker is on frequency.—Ed.) Since the nominal frequency of the crystal is 100.00 kHz, the measured frequency is within the quoted tolerances.

The oscillator provides an ideal frequency check for the Argonaut and other radios that lack an internal frequency marker. Statek manufactures the CX-1H crystals and provides circuit component values for frequencies from 10 to 600 kHz. Thus, most HF receiver calibration can be accomplished with this inexpensive and easy-to-build circuit. —*Michael C. Schell, KF8CZ, 7647 White Oak Dr, Solon, OH 44139*

[1]The CX-1H is available from Statek Corp., 512 N. Main St., Orange, CA 92668, tel 714-639-7810.

From January 1987 QST, p 27.

Some Power-Supply Design Basics

Part 15: Know your components and how to apply them correctly when designing a ham-shack power supply. Failures can be avoided and performance may be improved by observing some basic rules.

By Doug DeMaw, W1FB
ARRL Contributing Editor
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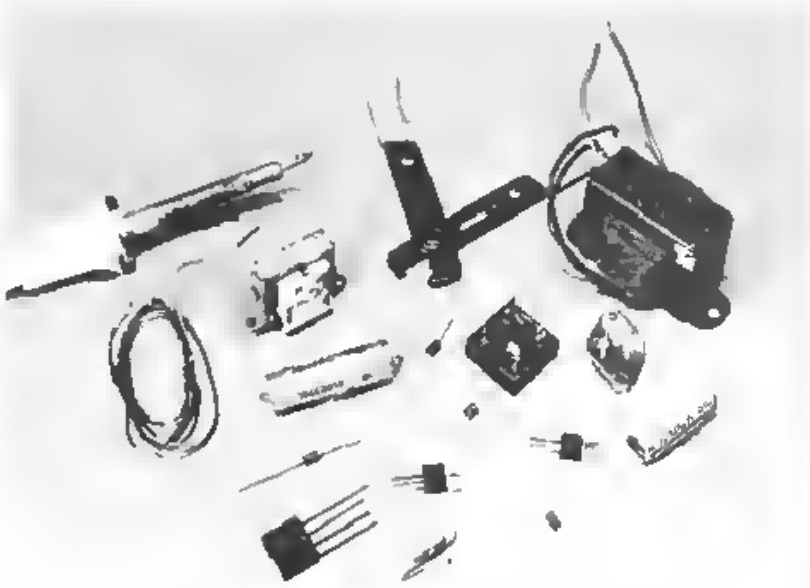

Let's think about power supplies in a practical way. I'll leave the exacting design information in the closet for this discussion. Those of you who want to dig deeper may consult the power-supply chapter of *The ARRL Handbook*, or such references as National Semiconductor's *Voltage Regulator Handbook* (1982). The latter publication contains several power-supply design application notes, along with countless equations for obtaining precise performance results.[1]

Rectifier Circuits

What are our choices for rectifier circuits, and what are the advantages and limitations of the various configurations? Fig 1 shows some of the possibilities we might consider. The most basic hookup we may use is shown at A of Fig 1. Here we have a half-wave rectifier with a single diode (D1) and filter capacitor (C1). The circuit simplicity is appealing, but regulation is very poor and the output ripple is high and hard to filter, compared to other circuits. Peak dc voltage across the diode may rise to 2.8 times the transformer secondary voltage (RMS) under no-load conditions with a capacitor filter. Conversely, the average output voltage, without filtering (under load) will be on the order of 0.45 times the T1 secondary voltage. The high no-load peak voltage, when filtered, results from C1 being charged. This stored voltage is then added to the peak voltage from the T1 secondary. These traits make the half-wave power supply suitable for low-current needs, such as bias supplies, but not for high-current applications.

A better scheme is shown at Fig 1B. Here we find the familiar full-wave rectifier. A center-tapped transformer is required, and the total secondary voltage must be twice that for a full-wave bridge circuit (C) for a specified dc-output voltage. The average output voltage from the diodes is 0.9 times half the RMS secondary voltage of T1. The peak output voltage (when using a capacitor-input filter, C1) is 1.4 times the T1 secondary voltage. Compared to the half-wave rectifier, this circuit requires less filtering because the output-pulse frequency is twice that of the half-wave rectifier. Also, each diode (D1 and D2) needs to accommodate only half the current taken by the load. This is because the diodes operate alternately at half cycles of the ac. The diode of circuit A must handle all of the load current.

Fig 1C shows the more common full-wave bridge rectifier. The principal advantage here is that no secondary center tap is required for T1. D1 and D3, in effect, provide the missing center tap. In this example two rectifiers operate on each half of the ac cycle. The average and peak output voltages for this circuit are the same as for the full-wave rectifier at B. The diodes should be rated for at least half the current taken by the load.

There are times when we need a plus and minus output voltage from a power supply. A simple method for obtaining the two equal voltages of opposite polarity is illustrated in Fig 1D. This dual-complementary rectifier requires a center-tapped transformer with twice the RMS secondary voltage of that for the full-wave circuit at C. You may think of this supply as two

[1] National Semiconductor Corp, 2900 Semiconductor Dr, Santa Clara, CA 95051.

For updated supplier addresses, see ARRL Parts Suppliers List in Chapter 2.

sections of the full wave circuit of Fig 1B. The notable difference is that two extra diodes are added (D3 and D4). They are connected for the polarity opposite that of D1 and D2. Peak and average dc output voltage is the same as that for circuits B and C of Fig 1.

Diode Selection

Earlier we discussed diodes that must handle one half the power supply load current. That is the *minimum* requirement. We need to consider peak currents when choosing our rectifiers. Using diodes that have marginal ratings for the intended application has caused many an amateur to scratch their head in wonderment after witnessing the failure of brand new replacement diodes in a repaired power supply! Be aware that the RMS current flowing into a capacitor-input filter is *two* to *three* times the dc output current. This is because the current is delivered in short pulses. A good rule of thumb (call it empirical if you wish!) is to use diodes rated at no less than twice the output current taken by the load. This allows ample leeway for the surge current of the power supply and has always provided reliability for me.

We must consider also the PIV (peak inverse voltage) or PRV (peak reverse voltage) of the diode we select. Earlier we learned that the peak voltage for a capacitor-input filter can rise to 2.8 times the RMS value of the transformer secondary winding. Therefore, our diodes should have a PRV rating of approximately three times the peak voltage value. If the transformer secondary RMS voltage is 12, the rectifier diodes should have a rating of 36 volts or greater. When building high-voltage power supplies, such as 2 kV, several 1000-PRV diodes must be connected in series in each leg of the rectifier in order to accommodate the high PRV. Equalizing resistors and capacitors are connected in parallel with each diode (as shown in the *The ARRL Handbook*) to equalize the voltage drop across each diode.

Choosing a Filter Capacitor

Amateurs tend to regard the filter capacitor as a casual matter. Why not simply use what is on hand in the junk box? Perhaps a randomly chosen capacitor value will provide adequate results, assuming output ripple is not a major consideration, and if the capacitor voltage rating happens to be sufficient. But wary of optimum performance? Well, there is a simple equation we may apply for low current power supplies when we are in doubt about the best type of capacitor to employ:

$$C_{\mu F} = \frac{I_L}{E_{P\text{-}P}} \times 6 \times 10^5 \qquad \text{(Eq 1)}$$

where I_L is the dc load current and E is the desired P-P output ripple voltage at 20 Hz. The P-P ripple voltage may be measured at the regulator output, under

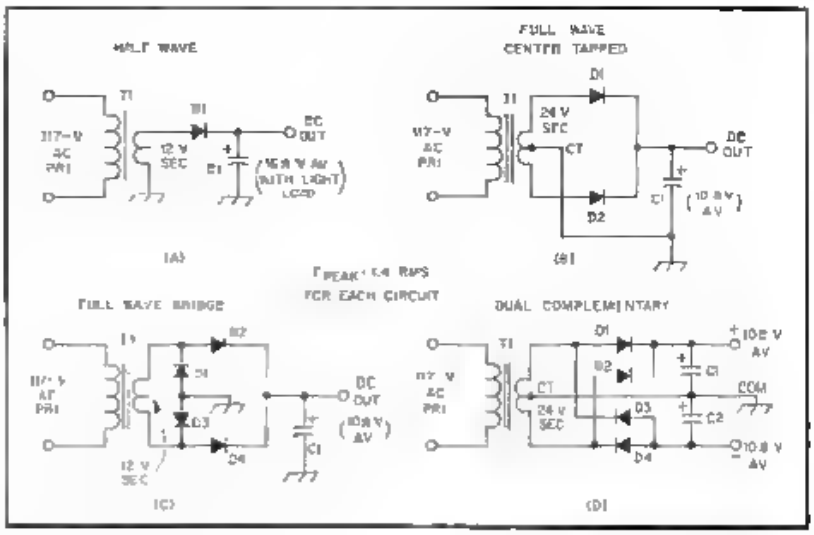

Fig 1—Conventional circuits for power-supply rectifiers. A discussion of these circuits is contained in the text.

normal current load conditions, with a scope.

Using Eq 1, we determine that a 3000-μF filter capacitor is required for a 1-A load current (12-V output), when the desired output ripple (under load) is 2 V P-P. Keep in mind that the values obtained from Eq 1 are based on the assumption that a regulator follows the filter capacitor. The regulator provides additional electronic filtering. The 3000-μF filter capacitor in the foregoing example should have a minimum rating of 36 V.

How about the Transformer?

A vital consideration when designing a power supply is that of the transformer rating—notably the secondary-current specification. Industrial design calls for some rather complex mathematical gymnastics, but we can follow a practical path when choosing the transformer we need for the job. Let's assume that we are using only the capacitor-input filter scheme, since it is more common and less expensive than the choke-input format. Based on this assumption our transformer secondary-current rating should be approximately 1.2 times the full-load dc current of the supply when using a full-wave center-tapped rectifier (Fig 1B). Thus, for a 2-A maximum load current the transformer secondary should have a minimum rating of 2.4 A (1.2 times the load current).

If we are using a full-wave bridge rectifier, the T1 secondary current minimum will be 3.6 A for a 2-A load (1.8 times the load current). Some amateurs have tried to use a 2-A transformer for a 2-A load, as an example, only to find that the transformer operated quite warm (even hot), and the output had substantial ripple under full load. If we take care in selecting our transformers, we will avoid these ailments. Make-do measures and junk-box components are not truly applicable when building a power supply.

Applying Regulators

Modern amateur equipment requires regulated dc operating voltage. The power supplies we have considered thus far are suitable for operating low-current devices or circuits that draw a steady current. When there are changes in load current, it becomes necessary to regulate the output voltage to ensure that the correct and safe operating voltage is present. Furthermore, the power supply should be relatively immune to momentary current overload and short circuiting. Present-day three-terminal regulator ICs offer the foregoing features. Many are capable of shutting themselves down when excessive current flows, which in turn protects the regulator, the attached equipment and the power-supply components.

Four basic considerations exist for selecting a regulator: (1) the maximum required output current, (2) required output voltage, (3) unregulated input voltage; and (4) ambient temperature. When you know the answers to items 1 and 2, you may consult the manufacturer's data sheets and make a device selection. Always choose a regulator that has a power dissipation (P_D) greater than the maximum load current presented by your equipment. Fig 2 shows some simple circuits for three-terminal regulators. The example at A represents a standard fixed-voltage regulator. ICs are available for various standard output voltages at various maximum-current ratings, such as 5 V, 8 V, and so on. They are also available for positive or negative power supplies. Fig 2B shows a typical adjustable regulator of the type that might

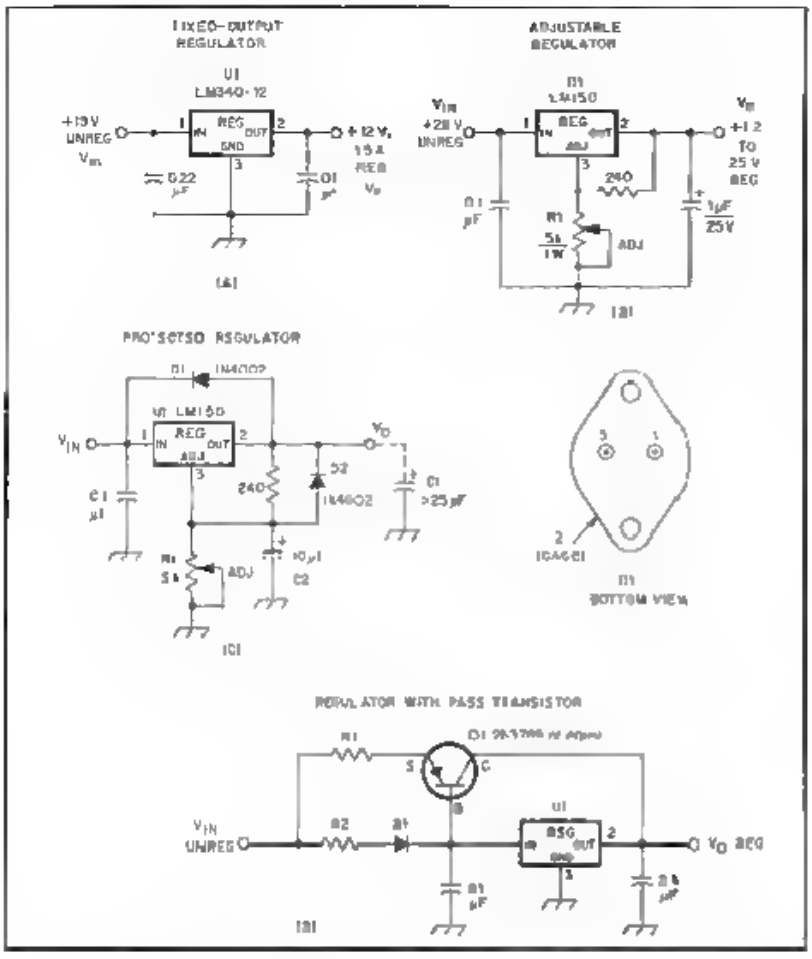

Fig 2—Circuits for three-terminal regulator ICs. These circuits are explained in the text.

be used for a bench supply. An adjustable regulator with protective diodes (D1 and D2) is shown at C of Fig 2. The diodes are recommended when the output capacitance (C1) is 25 µF or greater. This may be the situation when the equipment used with the power supply contains a high-value filter capacitor at the voltage-input terminal. The low internal resistance of the capacitor can cause high-amperage spikes when shorted (in excess of 20 A), and this can destroy the regulator IC. D1 protects U1 against input short circuiting (C1), and D2 protects U1 against output shorting (C2). Under the respective shorting conditions, C1 and C2 will discharge through the IC and destroy it.

We frequently need greater output current than a three-terminal regulator can provide. The solution to our problem is found in the circuit of Fig 2D. Q1 is a wrap-around pass transistor which handles the high current that U1 cannot accommodate. Several pass transistors may be used in parallel to increase the current rating of the regulated supply. Design information relating to this subject may be found in the 1987 edition of *The ARRL Handbook*, page 27-23.

At the start of this section we considered four items in selecting a regulator. No. 3 deals with the unregulated input voltage. Most manufacturers rate their regulators for maximum safe input voltage for fixed-voltage regulators that use ground as a reference. The maximum input-output voltage differential is used for adjustable regulators that do not use ground as a reference. This is sometimes specified as "input-output voltage differential." For example, Fig 2A has a "differential" of 7 V between pins 1 and 2 of the regulator, U1. The greater the input voltage, respective to the regulated output voltage, the higher the power dissipation within the regulator. Unnecessary power dissipation inside the IC requires greater heat sinking in order to keep the regulator within safe ratings. An example of wasted power and increased heat is seen when an input voltage of 25-28 is used for a 12-V regulated supply. A better input-voltage value is 18-19 V.

Item 4 relates to the ambient temperature of the regulator IC. This concerns item no. 3 and the size of the heat sink we employ. Thermal considerations represent a rather exact science that includes the junction temperature of the regulator. Another complex factor is the thermal resistance of the bond between the device and the heat sink. In any event, the regulator IC and the heat sink should never be more than comfortably warm to the touch after a period of full-load current flow. When in doubt, choose a heat sink that is larger than your intuition suggests. Be sure to use a *thin* layer of heat-sink compound (available at Radio Shack) between the regulator IC and the heat sink. The mounting screws should be snug but not too tight. Excessive torque may distort the IC and weaken the thermal bond, and it might even cause internal damage to the IC!

We must also be concerned about the operating temperature of the rectifier diodes. When large currents pass through the diodes, it becomes necessary to use heat sinks to keep the diodes cool. Bridge rectifier modules (four diodes encapsulated in a plastic block) lend themselves nicely to heat sinking. This is not true of plastic encased single diodes. Stud-mount, discrete-diodes are more suitable for use with a heat sink. In any event, the rectifier diodes, under full load, should never become hot to the touch.

A Practical Regulated Supply

Let's assemble the suggestions in this article and apply them in a small regulated supply that is aimed especially at the QRP operator. Fig 3 shows the circuit for a 1.5-A, 12-V regulated dc power supply. The component ratings are based on the guidelines given earlier.

Some additional parts appear in the diagram of Fig 3. They include C1-C4, inclusive, and RFC1. These units have been added to prevent unwanted common-mode hum in direct-conversion receivers. This malady is caused by RF energy from the receiver local oscillator (radiated by the antenna and power-supply leads) reaching the rectifier diodes. The RF energy is then modulated by 120-Hz energy and re-radiated. It is picked up by the antenna and heard as a hum in the DC receiver. C1-C4, inclusive, bypass the rectifier diodes at RF, thereby preventing them from acting as mixers or modulators. In effect, the capacitors provide an RF-current short across each diode.

As a further aid in solving the hum problem, we have included RFC1. It is a toroidal bifilar RF choke that prevents RF energy from entering the power supply via the power-supply leads. This preventive measure was first introduced by Wes Hayward, W7ZOI. I have found it to be effective with such transceivers as the HW-7 and HW-8. The bypass capacitors across the transformer primary winding

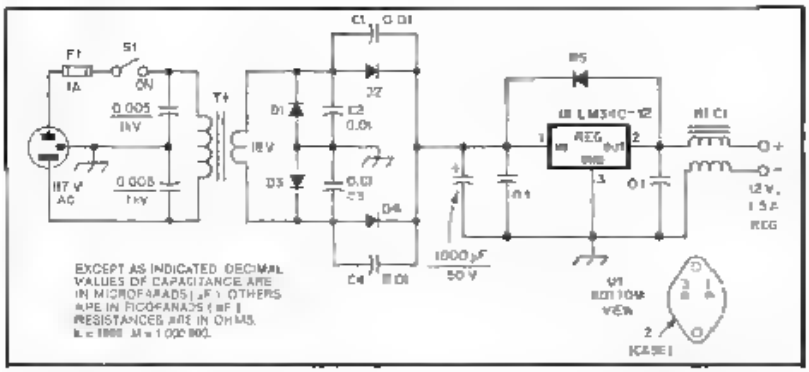

Fig 3—Schematic diagram of a practical 12-V, 1.5-A regulated dc power supply. Components C1-C4, incl, and RFC1 may be omitted if this circuit is not for use with a direct-conversion receiver (see text). Capacitors are disc ceramic except for the one with polarity marked, which is an electrolytic. All diodes are 3 A, 50 PRV or greater. RFC1 has 15 bifilar (parallel) turns of no. 14 enam wire on an Amidon Assoc FT-114-43 toroid core (1.14-inch-OD core with 850 μ). T1 has an 18-V secondary at 3 A or greater. U1 is a National Semiconductor Corp 12-V positive regulator in a TO-3 case. D1-D4, incl, may be a full-wave rectifier block (use heat sink here and on U1); see text.

also aid in keeping RF energy out of the power supply.

I recommend a modular bridge rectifier for D1-D4, inclusive. It should be mounted on a heat sink that is approximately 2 to 3 inches square. A 3-inch-square finned heat sink should be ample for U1.

Some Final Thoughts

We have merely agitated the surface in our basic look at power supplies. A lengthy book is needed to cover the subject properly. But, perhaps this article can provide some of the answers you have needed to fundamental questions about power supplies and the ratings of their component parts.

A number of regulated power supplies are described in detail in the 1987 edition of *The ARRL Handbook*. Additional design data may be found in the publication referenced in note 1.

From September 1989 QST, p 22.

A 1.25- to 25-V, 2.5-A Regulated Power Supply

Let's discuss the practical aspects of a test-bench power supply that's easy to build and get working. Most of the parts are available as surplus.

By Doug DeMaw, W1FB
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Luther, MI 49656

I needed a regulated 24-V power supply for development work with power FETs, but my lab supply could not deliver the current required because it provides a maximum of only 1.5 A. My work called for a current range from 2 to 2.5 A. Although I found a number of surplus fixed-voltage power supplies offered at modest prices, they were not variable-voltage units, and they qualified for the "boat anchor" weight class! I chose a typical amateur solution: build the power supply and make it compact.

This article covers the essentials of a simple power supply that you can duplicate in a few evenings. It can be expanded easily to deliver greater output current. The heart of this power supply is contained on a PC board that is available from FAR Circuits.[1] In fact, most components are available from mail-order houses.

Circuit Details

Fig 1 shows the circuit for my supply. The components marked with a double asterisk are external to the PC board. I recommend that you read the *ARRL Handbook* (1989 or other recent editions) for an explanation of how regulated power supplies operate. See pages 27-12 and 27-13 for a design description of a similar power-supply circuit.

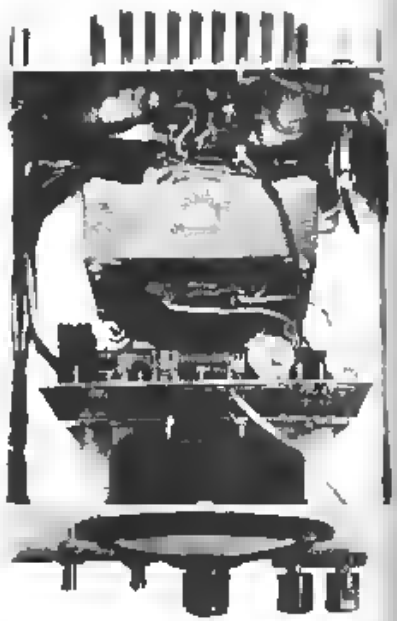
T1 is chosen for the voltage and current you require. You can use a 24-V transformer if you can work with a voltage range of 1.25 to 24. Select a transformer that can deliver 0.5 A or greater current than the maximum direct current you need. Likewise, use rectifier diodes that are rated for substantially more direct current than the supply will deliver. The PIV rating should be at least twice the secondary voltage of T1. U1 is a rectifier module that contains four 6-A, 200-PIV diodes in a full-wave bridge hookup. U1 is mounted on a small heat sink. I used a Thermalloy 6118B that is sold by BCD Electro.[2] The heat sink helps to keep the diodes from overheating when heavy current is flowing.

DS1 is a red LED that serves as the POWER ON indicator. You can replace the LED with a 28-V pilot lamp. If so, eliminate R10. By placing the LED or lamp in this part of the circuit, you will always know if the fuse, T1 and U1 are functional.

R1, R2 and R7 can be wound from no. 28 enamel wire on insulated forms, such as the body of a 10-kΩ, 1-W carbon resistor. You will need an accurate way to measure the wire resistance if you do this. These resistors are available from Mouser Electronics.[3]

U2 is a 1.25 to 30-V, 1.5-A three-terminal positive regulator. This device is also mounted on a small heat sink. I used a Thermalloy no. 6098 that I obtained from All Electronics Corp.[4] You can build your own heat sinks from 16-gauge aluminum or brass. Form U-shaped channels that are approximately 1-1/2 inches square by 5/8 inch high.

Q1 is a PNP (TO-204 case) power transistor. I recommend a Radio Shack® MJ2955 or RCA SK3335 transistor. These have a 150-W rating. The emitter and base pins are bypassed to ground at the pins by means of C7 and C8 in Fig 1. This is a preventive measure against instability, owing to the long leads between Q1 and the PC board. You can parallel two or more pass transistors to increase the output current of the supply. Each pass transistor provides an output-current increase of approximately four times that of U2. The single device at Q1 in Fig 1 ensures an out-



Internal view of the assembled power supply. The chassis and panels are made from single-sided PC board. The circuit board is mounted vertically to conserve space.

[Notes appear at end of article.]

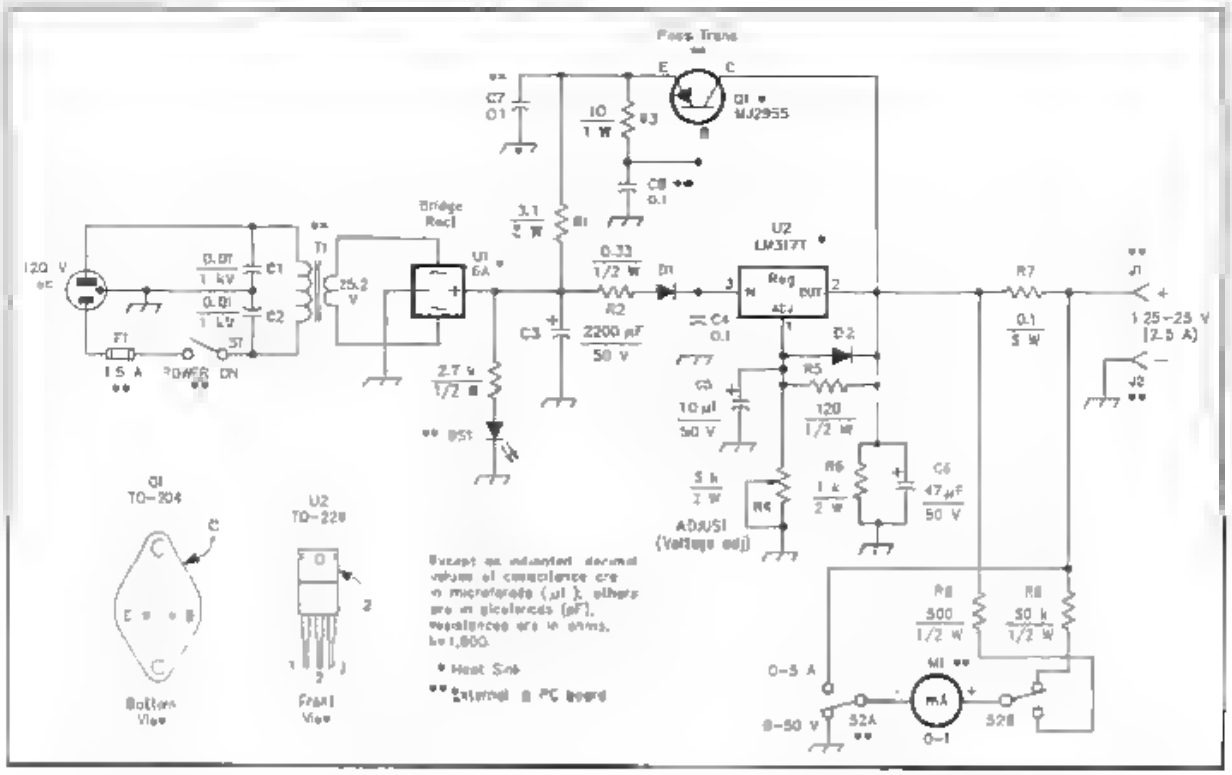


Fig 1—Schematic diagram of the 1.25- to 25-V regulated power supply. Capacitors are disc ceramic except for those with polarity marked, which are electrolytic. See text for data concerning heat sinks for Q1, U2 and U3.

D1, D2—1-A, 100-PIV rectifier diode.
DS1—Red LED.
F1—1.5-A 3AG fuse in chassis-mount holder.
J1, J2—Standard five-way binding post, one red, one black.
M1—Milliammeter 0-1 mA dc (see Notes 5 and 9).
Q1—PNP power transistor MJ2955 (Radio Shack) or equiv device with a +70-V, 10-A, 150-W rating in a TO-204 case.
R1, R2, R7—5-W wire-wound resistor. See Notes 3 and 4 for source. Q1, use 17 inches of no. 26 enam wire, single-layer wound, on a 10-kΩ, 1-W carbon-composition resistor for R1 and R7. For R2, use 26 inches of no. 30 enam wire on a 10-kΩ, 1-W carbon-composition resistor (scramble wound).
R4—Panel-mount, 5-kΩ, 2-W or 5-W potentiometer, carbon or wire wound (see Note 8).
R8, R9—See text.
S1—SPST toggle switch.
S2—DPDT toggle or rotary wafer switch.
T1—25.2-V, 2.75-A power transformer (see text).
U1—6-A, 200 PIV bridge rectifier with heat sink. See text.
U2—LM317T +1.25- to 30-V, 1.5-A TO-220 regulator. Use an LM317HVK (TO-204 case) for dc output voltage greater than 40. See text.

put current of 5 to 6 A if the transistor has a large enough heat sink to remain at a safe operating temperature. If you use additional pass transistors, you will need to replace T1 with a heftier transformer.

Output voltage and current monitoring is done with a 0-1 mA meter (M1). I used a surplus meter I had available, hence the additional scales on the meter face. A suitable 2½- or 2-inch meter can be purchased from Dick Smith Electronics.[5] The voltage drop across R7 indicates the current being taken by the load. R8 allows M1 to read 0.5 V full scale, which corresponds to 5 A of current through R7. R9 permits the meter to read 50 V full scale. Try to use 1% resistors for R7, R8 and R9 for best meter accuracy. I used two 1-kΩ, ¼-W resistors (5% tolerance) in parallel for R8 and two 100-kΩ, ¼-W resistors in parallel at R9. R7 is my unit as a 5% resistor. The accuracy of the readings is satisfactory for my work.

You can lift J2 above common ground if you want to extract negative voltages from the power supply. A third binding post can be added (common to the chassis) for connection to J1 or J2, depending on the desired polarity. If this is done it will be necessary to bring all of the negative circuit leads to a bus that connects to J2, except for C1, C2, C7 and C8.

Construction Notes

The photograph shows the interior of my power supply. I used an old cabinet that a welder friend had made for me some 25 years ago. The chassis and panels are made from single-sided PC-board material (metal side in). The mating surfaces are soldered together. I used gray automotive primer as the undercoating for the cabinet, then sprayed it with clear lacquer. The panel has gray primer for the undercoating and white spray enamel as the finish coat. Clear lacquer was sprayed over the white panel after the decals were added. The cabinet dimensions are (HWD) 6 × 6 × 8 inches.

You can see in the photograph that the PC board is mounted vertically to save space. It is held in place by an L-shaped aluminum bracket. Q1 and its heat sink are attached to the rear outer wall of the chassis assembly. My heat sink is a surplus extruded type, measuring 3¼ × 3¾ × 1 inch. I do not recommend a Q1 heat sink that is smaller than 13 square inches by 1 inch thick. Larger heat sinks will provide added Q1 protection. A hefty heat sink is available from Dick Smith Electronics (no. DS-H3471).[6] The photograph shows a thick heat sink with fingers. It was replaced by a heavier, extruded unit of the type just mentioned, owing to excessive Q1 heat during high-current periods. John Meshna Jr, Inc lists a dual TO-3 (TO-204) heat sink (no. SP-58A-28) that is suitable for one or two pass transistors.[7]

You may find that R4 and R6 are difficult to locate. Wire-wound or high-wattage carbon potentiometers are scarce items on the surplus market. I was able to find a 2-W

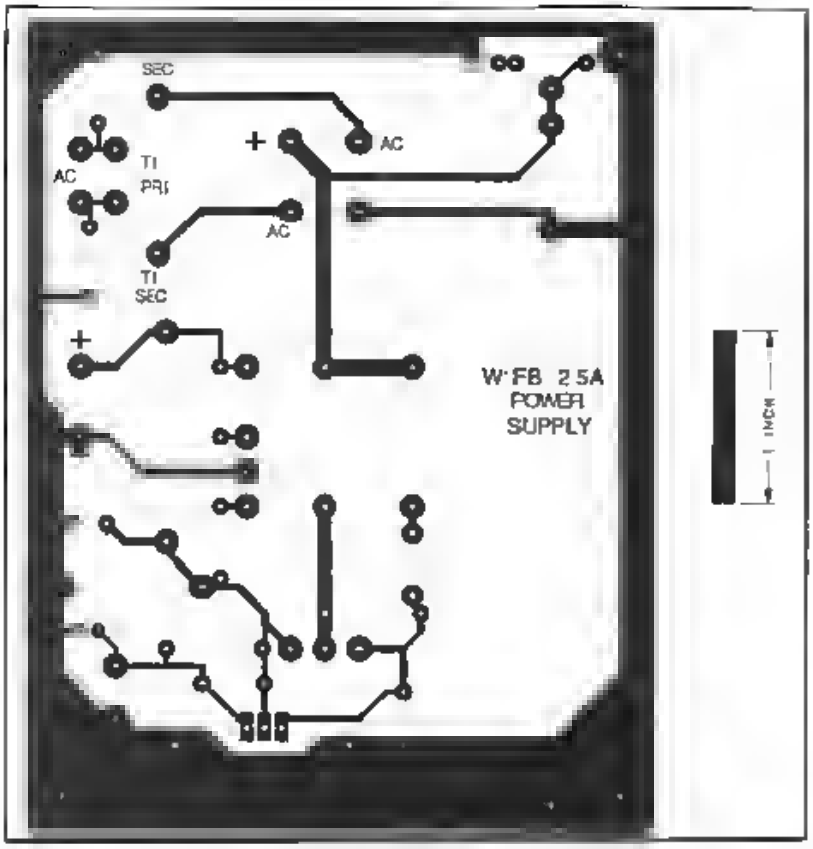


Fig 2—Circuit-board etching pattern for the power supply. The pattern is shown full size from the foil side of the board. Black areas represent unetched copper.

much tension causes stress that can damage the semiconductors.

Use 16- or 18-gauge insulated hookup wire between the T1 secondary and the PC board, and likewise between J1 and the PC board. This will minimize unwanted voltage drops through these wires. Also, use insulating hardware to isolate Q1 and U2 from their heat sinks, unless the sinks are "floated" above chassis ground. Radio Shack has insulating kits (no. 276-1371 for Q1 and 276-1373 for U2).

A scale PC-board etching pattern is shown in Fig 2. A parts-placement guide is provided in Fig 3 (see Note 1).

Summary

Many hams have told me they don't build equipment because "it's impossible to find the parts." Perhaps the references

Fig 3—Parts-placement guide for the circuit board, not to scale. Parts are placed on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. Component outlines are not necessarily representative of the shapes of the actual parts used.

(5-kΩ) control in the Jameco catalog (no. CMU-5021).[5] It is a chore to locate 2-W carbon resistors. If you can't find the proper unit for R6 of Fig 1, you can parallel two 2.2-kΩ, 1-W resistors.

As mentioned earlier, most of the parts for this project can be purchased by mail. The LM317T, for example, is available from the suppliers listed in Notes 2, 4 and 5. U1 can be purchased from BCD Electro (see Note 2) or from Mouser Electronics (no. 33BR062—see Note 3). C3 can also be obtained from Mouser (no. 20NR905). I purchased T1 from Electronic Surplus, Inc (no. 767B11).[9] If you desire an output voltage greater than 25, you can buy a 32-V, 3.5-A transformer from Fair Radio Sales (no. X5157308).[10] The increased dc voltage (46 V maximum) will require that you replace U2 of Fig 1 with an LM317HVK, which is supplied in a TO-204 case. The use of this IC requires a modification of the PC board in Fig 2.

You can buy a modestly priced 0-1 mA dc meter from Fair Radio Sales, which offers a 3½-inch round unit that has a 0-50 scale (ideal for this project). The cost is $5 at this writing.

Be sure to use a thin layer of heat-sink compound or silicone grease between Q1, U1 and U2 and their respective heat sinks. Affix the three devices firmly (but not excessively tight) to the heat sinks. Too

in this article will make your job easier—and they should also be useful when searching for parts to use in other projects.

The maximum recommended load current versus output voltage for the circuit in Fig 1 is 500 mA (1.5 V), 750 mA (6 V), 1 A (9 V), 1.5 A (12 V), 1.75 A (18 V), 2 A (20 V) and 2.5 A (25 V). These figures are for steady-state load current. For intermittent loads, such as for CW and SSB transmitters, the current maximum can be increased 25 to 30 percent, assuming a typical duty cycle during transmit.

This power supply is certainly suitable for uses other than a test-bench unit. It can be used to operate a low-power VHF transceiver or homemade QRP gear, or as a battery charger. Good luck and have fun!

Notes

[1]FAR Circuits, 18N640 Field Ct, Dundee, IL 60118, tel 312-426-8451, evenings. Price: $8.50 (includes shipping to US addresses).
[2]PO Box 830119, Richardson, TX 75083-0119, tel 214-343-1770 (catalog available).
[3]Mouser Electronics, PO Box 699, Mansfield, TX 76063, tel 800-346-6873 (catalog available).
[4]All Electronics Corp, PO Box 567, Van Nuys, CA 91408, tel 800-826-5432 (catalog available).
[5]Dick Smith Electronics, PO Box 468, Greenwood, IN 46142, tel 317-888-7265 (catalog available).
[6]See Note 5.
[7]19 Allerton St, Lynn, MA 01904, tel 617-595-2275 (catalog available).
[8]Jameco® Electronics, 1355 Shoreway Rd, Belmont, CA 94002, tel 415-592-8121 (catalog available).
[9]Electronic Surplus, Inc (formerly R&D Electronics), 1224 Prospect Ave, Cleveland, OH 44115 tel 216-621-1052.
[10]Fair Radio Sales Co, PO Box 1105, 1016 Eureka St, Lima, OH 45802, tel 419-227-6573 (catalog available).

For updated supplier addresses see ARRL Parts Suppliers List in Chapter 2.

From September 1987 QST, p 17.

Alternative Energy—An Overview of Options and Requirements

Part 1: Planning on operating far from the power grid? You *can* have the electrical energy you need when you need it, but it takes a systems approach. Here's a look at how to pull energy from sun, water and wind.

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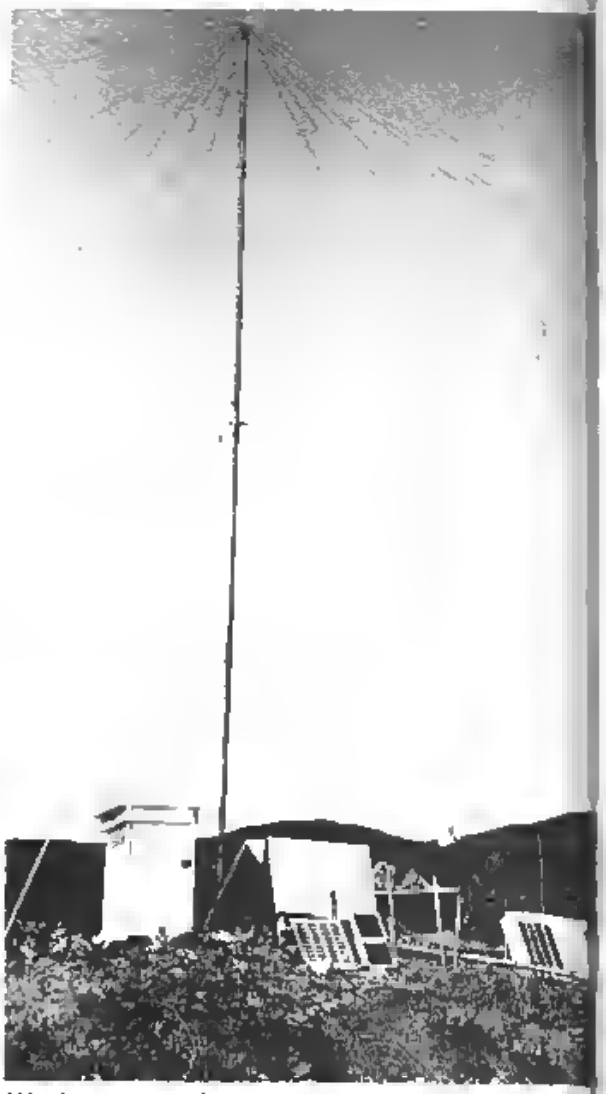


Miles from commercial electric energy, Mike Miceke's experimental 1750 meter beacon has chugged along on solar power since 1984. The antenna is an 18-foot vertical. *(photos by WB6EER)*

Radio amateurs have always found many reasons to operate beyond the power lines. Field Day, DXpeditions, mountaintopping for the sheer fun of it, emergency work from disaster sites where power is out, and from wildfires in places where power lines have never run—all of these situations call for portable and more or less independent energy resources. Some amateurs find themselves spending long periods in locations far beyond the reach of commercial energy distribution. Others need to operate repeaters or remote equipment in places where commercial energy is either unavailable or unreliable. Still others find themselves caught up in the challenge of developing their own energy resources although they have perfectly good ac available in their wall sockets.

I suspect that the term "alternative energy" evokes quite a variety of responses and definitions in the minds of *QST* readers, so before moving into the subject proper, I'll discuss my personal definition of the term and where I stand in relation to this exciting field.

In my opinion, energy alternatives are those that provide electrical or other energy in some fashion not directly connected to commercial generation and distribution networks. If you buy a gasoline powered generator and fuel to operate a Field Day rig, I think it is stretching things a bit to say that you're using alternative energy. If, however, you modify the generator to operate on methane, alcohol or wood, and then proceed to produce the fuel before going on the air, *then* you're on alternative power!

A combination of random circumstances moved me beyond the reach of power and telephone lines in 1969. Somehow, I have never gotten back to "civilization," except as a visitor. An addiction to personal comfort, late night reading and a variety of technical hobbies all combined to motivate my alternative energy efforts, so I now find myself operating on a mixture of solar, hydroelectric and internal-combustion derived electricity. My commercial energy source is a 2.5-kW Onan power plant attached to a 250-gallon propane tank. The tank is topped off once a year. This generator sees limited duty, operating a washing machine once a week, a 16-mm movie projector on rare occasions, and power tools once in a while, as needed.

The combined solar and hydroelectric operations provide power for lights, amateur and experimental radio stations, soldering irons, an electric typewriter and other apparatus. The economics of our situation dictate a piecemeal approach with maximum emphasis on scrounging, salvage and modification of available devices. My family and I know we'll find plenty of uses for all the energy we can afford to produce or store. At the same time, however, we organize our activities around whatever energy happens to be available at any given moment. Present solar capacity at the homesite is 90 W peak. Maximum hydroelectric capacity is around 300 W. Full hydroelectric potential is generally available from December through

May, while solar input is best from May through August. Some conservation measures are generally required from September through mid-November.

Are we self-sufficient? By no means! Full energy self-sufficiency is not a realistic concept as applied to alternative energy users. No matter how we generate and store electricity, we are ultimately dependent upon energy-intensive technologies to provide our essential hardware. Nor does alternative energy mean free energy. In most cases, cost per watt-hour (Wh) is substantially higher than commercial power. The exceptions are efficient systems that can operate for a long time (decades) at minimal cost above the initial installation expenses.

The effort necessary to achieve any degree of energy self-sufficiency almost inevitably leads to some realization of how much power a person or family really *needs* in order to live comfortably; generally, quite a bit less than the average use of "cheap" commercial power consumes. Thus, even in the short term, a small quantity of expensive power may be just as satisfactory as a lot of cheap power: Wise use of a scarce resource can be as effective as luxurious use of an abundant resource.

Achieving energy self-sufficiency engenders an awareness of the degree to which the global population is becoming ever more dependent upon the electrical power network. This potent but fragile network is, to a distressing degree, taken for granted by its users. When we build our own power systems, we come to better appreciate both our needs for electrical energy and the awesome scope and complexity of the commercial systems that are a cornerstone of modern civilization.

Short- or Long-Term Energy Needs?

If alternative energy signifies independence from the mains, *how long* must that independence be maintained? Will an alternative energy system serve only during emergencies and self-initiated portable/mobile stints, or will it be part of your way of life—perhaps permanently?

With NiCd battery packs and chargers available almost everywhere, much portable Amateur Radio operation can almost be considered an extension of the power lines. Potentially, if not always in fact, portable work goes well beyond this simple view. Portable operation, no matter how casual, requires energy storage and management. Such techniques are basic to nearly all alternative energy systems. Secondly, there are many alternatives as to how we charge those batteries. When there is no electricity in the power line or when there is no power line, how do you keep that hand-held "transceiver running? Answer: with energy from an alternative source.

The most rigorous applications of alternative energy techniques involve permanent or semipermanent installations that must, for whatever reason, be energy self-sufficient to some degree. Such installations, be they repeaters or small households, must be dependably supplied with energy despite the fact that there can be no expectation of regular reconnection to the mains. For the remainder of this article, we'll be exploring alternative energy systems that can meet these needs. Energy for the short-term needs of portable, mobile and emergency Amateur Radio operation has received ongoing coverage in *QST* and *The ARRL Handbook*. Portable and mobile ham operation *do* have something in common with long-term alternative energy production, however. See the sidebar, "Milking a Herd of Cars."

Milking a Herd of Cars

Many of us have participated in mobile Amateur Radio operation, mainly from vehicles using 12-V electrical systems. Most vehicular electrical systems have sufficient capacity to operate a wide range of solid-state equipment with little or no modification to the power source. Does this mean that the nearest car or truck is an alternative energy source just waiting to be tapped?

Whether we consider vehicles to be a valuable source of alternative energy depends somewhat upon how we look at them. Viewed as a *resource*, motor vehicles are ubiquitous. They have on-board energy production, storage and regulation systems. Viewed purely as an electrical energy source, they are dreadfully *inefficient*. Even so, in certain types of emergencies, motor vehicles may provide the best (or only) short-term source of electricity. Further, the family car can be milked for a fair amount of battery charging, though at some reduction in gas mileage—not a particularly efficient source of energy, but a source nonetheless.

Automotive electrical systems are valuable in alternative energy production for another reason: Their components may be used in *long-term* energy production. Automotive storage batteries (and their relatives) may be the first such components that come to mind, but the list doesn't stop there. In my small hydroelectric installation, I get thousands of hours of service from used alternators purchased cheaply in scrapyards. They don't last forever, but with a typical life of three to four years, the annual cost of one alternator is under $5.

Energy Production

Most small-scale alternative energy schemes likely to be undertaken by radio amateurs involve two distinct processes: production and storage. It's easier to talk about storage if we have something to store, so let's look first at means of producing electricity.

Wherever there is an energy gradient, be it mechanical, chemical or thermal, there is the possibility of producing electricity.

Wherever there is an energy gradient, be it mechanical, chemical or thermal, there is the possibility of producing electricity. Wherever there is motion, kinetic energy may be intercepted and put to work, either directly to drive machinery or indirectly to generate electricity. The classic and most readily exploited energy sources are water and wind. Where electricity is the desired product, the linear movement of water or wind is transformed into a rotary movement suitable for driving generators. This transformation of linear to rotary motion is usually accomplished with a fan-like propeller in the case of wind, and any of a variety of turbines and wheels in the case of water.

Wind and waterpower systems offer an interesting analog to electrical theory. We may extract equal amounts of energy from large volumes moving at low velocity or small volumes moving at high velocity—very much as 1 W can be developed with 1 V at 1 A or with 1 kV at 1 mA. In practice, water is more easily manipulated in this way, while wind must be taken as it comes. Still, in areas where winds are light, larger wind turbines may be used to generate the same output produced by smaller generators in windier areas.

Waterpower Systems

Available water pressure—usually the result of water running downhill—is the heart of every hydroelectric energy system. Pressure increases in direct proportion to the height of the water column contained in the supply pipe or *penstock*. The height of the water column is defined as the *head*. Pressure at the bottom of the penstock in lb/in^2 will be approximately half the measure of the head in feet. Friction must be considered as well, because some of the energy in moving water is spent in friction between water and piping. This factor depends on pipe length and diameter, as well as the pipe material and its condition.

Hydroelectric systems can be divided into high- and low-head categories. In high-head systems, energy is extracted from a relatively small volume of water moving at high velocity, while large volumes of slow-moving water are employed in low-head operations. Either approach may be used to produce a few watts or many kilowatts

of power, and the appropriate choice will depend largely on the nature of the available water supply.

Low-head systems entail construction techniques suitable to routing and controlling large volumes of water. If a head of 15 to 20 feet is available, with flow enough to fill an 8-inch (or larger) pipe, turbines and complete generating systems are available that will produce ample electricity to maintain a household, assuming that energy *needs* versus wants have been carefully assessed. Despite the considerable volume of water they entail, low-head systems can be relatively compact because the water source usually need not be far from the generator. Low-head systems based on wooden waterwheels, such as those found at water-powered mills, can be built from basic materials with a minimum of precision work. Although such wheels can deliver substantial power, they do not turn fast enough to drive generators directly, so their speed must be stepped up mechanically. Because the required step-up is usually too extreme to be accomplished efficiently with belt-and-pulley drives, the usual solution is gearing. Automotive gear trains—even whole transmissions—are a common choice for this gearing.

High-head systems require less water for a given output, so they can be applied in areas where water volume is insufficient for low-head generation. A complete high-head installation may require a lot of territory. Much pipe may be needed. Safe and reliable containment of high-pressure water may dictate the use of expensive construction materials. A variety of small turbines, mostly of the Pelton type, are available for high-head systems. Turbine diameter may be as small as four inches for very small systems, with diameters of eight to twelve inches common in 1- to 3-kW applications.

For optimum efficiency, a turbine must be matched to both the volume and velocity of available water. Nonetheless, turbines can deliver useful power over a wide range of pressure and flow, so if you come into possession of an old unit that isn't exactly right for the available situation, it may still be worth using.

Hydroelectric systems may be designed to deliver either low-voltage dc or 117 V (and higher) ac, depending upon the power available and your requirements. Even a tiny hydroelectric potential of 2 or 3 A at low voltage can be useful in battery charging service. Because such a charger can work 24 hours a day, its performance can approach that of a photovoltaic solar-powered) unit having four times the output capacity.

The construction of any hydroelectric system represents a long-term investment. Careful planning is a must. Attention must be given to the development of an adequate water source, disposal of water discharged from the system and a multitude of mechanical and environmental concerns related to the routing of water. Potential impact on erosion and vegetation patterns should be considered carefully. A good hydroelectric installation should deliver its design capacity for many decades while requiring only a relatively low level of maintenance and post-construction expense. Of the available small-scale alternative energy techniques, only hydroelectric systems offer a continuous supply of electricity with no battery storage requirement.

Fig 1—Hydroelectric power, anyone? Here, the author's 9-inch Pelton turbine (bottom), manufactured circa 1890, drives a modified automotive alternator (upper left) to produce 117 V ac. Pressing the button at the upper right of the meter panel provides dc field current for the alternator at system startup.

In my own system, I use a nine-inch Pelton wheel (manufactured circa 1890) with a 180-foot head delivered via ¼ mile of 2¼-inch pipe. The pipe is installed in six-foot sections, and I'm sure that friction losses are high in this piping because of turbulence at its many joints. Why the short pipe sections? They were brought to the generating site in the 1930s—packed on mules! In the original installation, the nine-inch turbine was used to operate circular and crosscut saws for cutting firewood. At the same time, a considerably smaller turbine charged batteries. Somewhere along the line, the smaller unit vanished. Since we can now take chain saws to the trees instead of dragging whole trees to the saw, the larger turbine is free for use in electricity production. The Pelton wheel drives a modified automotive alternator through a belt and pulleys (see Fig 1). In my

case, the alternator modification is a simple matter of installing wires to bypass the connection of one of its three poles to rectifier diodes so that low-voltage ac is available. This ac is applied to the primary of a 12-to-117-V step-up transformer, providing up to 300 W of ac power for lighting, soldering and small electronic applications. Operating frequency and output voltage in this system depend somewhat on loading—the resulting electricity is definitely *not* a 60-Hz sine wave!—but the energy generated is highly useful nonetheless.

Such a simple system is feasible only because of the limited power available from the turbine. Higher power would call for closer regulation and full utilization of the alternator's three-phase capacity.

At times, I use the waterwheel to charge batteries. To do this, it is necessary only to switch the transformer out of the alternator circuit and connect a battery to the alternator's dc output. In this mode, the alternator functions just as it does in a car except that output is somewhat reduced because of the limited driving power. If I want a trickle charge, I just close the turbine water supply valve until I get the charging current I want.

Wind Power Systems

As with water power, the kinetic energy of wind may be tapped by converting its action to a rotary motion suitable for operating generators. Like water, wind can provide large or small amounts of power, depending on need and subject to prevailing conditions. But unlike water supplies, which can be closely regulated for continuous operation, the wind is always somewhat intermittent and highly variable in intensity.

Small-scale wind power generators are generally employed as battery chargers, the stored energy being used directly or after conversion to ac for the operation of 117-V appliances and tools. I have heard that on the coast of Chile, where evening wind is common and fairly constant, extensive use has been made of propeller-driven automotive generators connected directly to low-voltage lamps.

The variable nature of wind and the extremes that will occasionally be encountered at any generation site necessitate mechanical means to prevent self-destruction of generating apparatus during high wind conditions. Such velocity compensation may be accomplished with variable pitch propellers that reduce efficiency as speed rises. Alternatively, tail-vane action may be used to turn the wind turbine off-axis to high winds. Various centrifugally actuated braking devices may be used; these operate aerodynamically or mechanically.

Provision must be made to maintain generation no matter what the wind direction. A tail vane will take care of this. Of course, if the generator is free to rotate as the wind turbine aligns itself with changing wind direction, a slip-ring coupling arrangement must be employed to get electricity out to the load. Voltage regulation is essential in wind systems to prevent overcharging of batteries and damage to low-voltage apparatus.

The preceding comments apply mostly to propeller-driven upright wind turbines common to most windmills. If a wind turbine is built similar to a vertical squirrel-cage fan, only the turbine itself need rotate, because such a turbine responds to winds from any direction. Thus, the generator may be fixed, and no slip rings are required to transmit its output to the load. An added advantage claimed for some squirrel-cage wind turbine designs is that they are self-regulating with respect to speed: Increasing wind velocity increases turbulence around the rotor effectively limiting its maximum speed.

In the days before rural electrification, many homes on the Great Plains were wind powered. A number of companies produced wind generators and complete power systems. As power lines spread through the countryside, wind power installations were shut down and largely left to deteriorate. During the 1970s, increasing interest in alternative power sources led to the salvage and reconditioning of many such units. Some usable salvage may still be gleaned from such systems, but unless wind equipment has been regularly maintained or carefully stored, extensive restoration is required. Any towers and tower-mounted units that have been neglected for decades are potentially dangerous. Undertake salvage efforts only with the help of properly equipped people experienced in wind-power machinery.

Recently, we've seen significant advances in wind power design; high-efficiency units with long lifetimes are available at a variety of power levels. For the dedicated home builder, much has been published covering wind power at all levels, from the most basic on up. These are too numerous to list here; your library is the best place to begin research. If you possess good workshop skills, study of a cross section of this material should enable you to set about designing and building your own wind power system.

Photovoltaic (Solar) Power Systems

Until recently, practical electrical generation has been either electrochemical (primary cells) or electromechanical (generators). With the development of photovoltaic (PV) technology, we are presented with a third and highly elegant option, the direct conversion of light energy to electricity. Modern PV devices can do this efficiently enough to power a wide range of electrical and electronic devices.

The production of PV materials is an energy-intensive process, but the practical application of PV products is the most straightforward of any alternative energy technique. Aside from switches and relays, PV systems entail no moving parts bigger than electrons and photons. Thus, maintenance requirements are minimal. PV generating panels should be kept reasonably clean because sunlight must be able to reach them, electrical connections must be sound, and batteries must be maintained in good condition.

Electromechanical systems have definite ranges of optimum efficiency. A system designed to serve a maximum load demand may not be "happy" with an average or minimum load. With PV technology, power is available in direct proportion to collector surface area, so system capacity may be easily tailored to specific load requirements by adding or subtracting collectors.

PV collectors are a long-term energy source. Most good-quality solar panels are guaranteed for five to ten years, but the usable lifetime of a panel will ordinarily extend for a considerably longer period, offsetting high initial costs with low maintenance expense and high life expectancy.

At the moment, PV efficiency per unit area appears to be rising faster than cost per unit area. Various long-awaited breakthroughs in manufacturing processes, especially the continuous roll-to-roll production of amorphous silicon alloys, are on the point of bringing the PV industry to a new and highly competitive level. *QST* recently carried a New Product announcement about amorphous silicon panels.[1] Although photovoltaics are an expensive energy source, they are indispensable and cost-effective tools when it comes to powering permanent and semipermanent remote installations such as repeaters, remote bases, beacons and sensors.

For powering isolated households, PV will likely be less cost effective than hydroelectric or wind systems, assuming that those resources are readily available for development. But this is really a complex question. Its answer depends on how much energy you choose to define as "enough," how peak requirements relate to average demand, and so forth.

Modern photovoltaic devices can convert light energy to electricity efficiently enough to power a wide range of electrical and electronic devices.

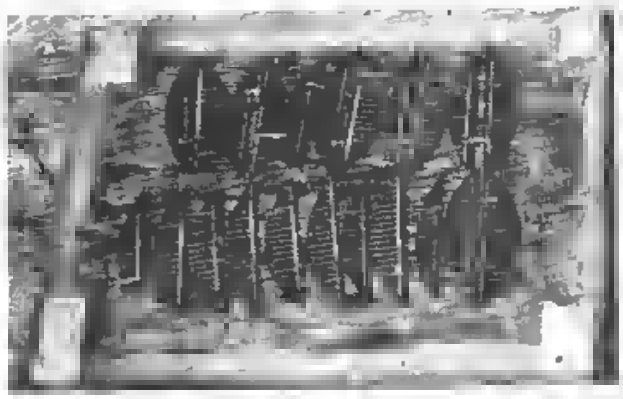
Often, no single energy source will readily satisfy the full range of load demands, so system combinations become very attractive. For instance, photovoltaic battery-charging capability can work in conjunction with an internal-combustion-engine-powered generator. During peak-

[1]Bruce O. Williams, "Sovonics Amorphous Solar Electric Panels," New Products, *QST*, May 1987, p 20.

Used Solar Cells Deserve a Place in the Sun

Several years ago, I constructed a small (12 V at 30 mA) energy panel from small semicircular solar cell scraps. The panel worked but it was never really weather tight. Moisture entered during every rain and heavy fog. This condensed on the glass surface of the panel, reducing its output until evaporation cleared the problem. One day, the panel took a fall and its glass cover broke. It lay abandoned and exposed to the weather for about two years until my curiosity led me to bring it home and check its cells. To my surprise, most of them performed quite well. Because a few cells were broken, and some soldered joints were in bad shape, I dismantled the panel, thinking that I might be able to salvage enough cells to charge small 6-V batteries. Before I could begin work on this charger...the PV cells, lying in disarray on a table, were accidentally exposed to the discharge of a Tesla coil. Sparks jumped freely between all of them! This looked like the end of my mini-project—but the cells checked out fine under test. Using them, I built the 6-V panel shown in the photo. For more than a year, it has been delivering 30 mA to my 6-V batteries whenever the sun shines.

I don't recommend trying any of these tortures on your solar cells, but my experiences do show that unlike many semiconductor devices, solar cells can survive abuse. Because of this, used solar cells should not be overlooked as a source of cheap energy.



This 6- x 8-inch solar panel delivers 30 mA to a 6-V battery under full sun conditions. The cover is acrylic plastic sheet, the back is plywood and the sealant is silicone rubber. These cells still work even after a rather shocking experience.

sun periods, the PV system maintains battery charge to operate lighting and low-voltage dc devices. At such times, the engine-driven system sees occasional use in running high-powered tools and appliances. When PV output falls because of short daylight hours in winter or extended periods of overcast, the engine system automatically tops off the battery charge. If PV capacity is later increased, the engine generator runs less, extending its life and reducing the fuel bill.

In many respects, photovoltaics are a nearly perfect power source, but there are drawbacks. System cost is extremely high if any attempt is made to satisfy the energy requirements of a typical American household with a PV system. Individual solar panels are light and fragile, although fragility stands to be greatly reduced with amorphous silicon PV collectors. They must be exposed in plain sight, creating a risk of theft and vandalism so great as to make PV installations totally impractical in many situations.

Battery storage is an essential adjunct to PV power systems. Many installations will also require the conversion of stored dc to 117 V ac, usually by means of high efficiency solid-state inverters. Outlays for energy storage and conversion hardware must also include the cost of suitable housing for the components. This calls for careful planning, which should begin with a realistic assessment of system requirements.

Many people look at the high cost of the solar panels and decide to build their own, either from kits or by using "bargain" cells available from many sources. With care, anyone can solder cells together and house them well enough to provide protection from the immediate effects of the weather—but by the time the materials have been obtained and the hours have been invested in construction, the "bargain" may not seem like such a bargain. The real question is how long such collectors will continue to deliver their initial output—or work at all, for that matter. The answer depends to a large degree upon the materials used and the care exercised in building the panels. Users looking for more than a few watts should be prepared to tackle a large, demanding (and expensive) construction job. Otherwise, purchasing guaranteed, ready-made units is more satisfactory. On the other hand, construction of a small PV panel can provide a good introduction to photovoltaic technology while adding a useful energy source to the shack.

The present generation of solar cells (of crystalline structure, as opposed to amorphous cells) is fragile. The cells are subject to outright breakage and to microscopic cracking, which interrupts conductive paths, reducing and eventually destroying cell capacity. Even so, solar cells can survive long use and even abuse, so used PV devices may be worth investigating. (See the sidebar, "Used Solar Cells Deserve a Place in the Sun.") As existing solar panels age and deteriorate, they are replaced by newer PV technology. This means that more and more surplus, defective and broken early-generation solar panels are sure to become available. These may well provide a useful source of materials for persons desiring to build their own panels on a low budget. While construction of a large panel means a major construction effort, small units are relatively quick and easy to build—and if the price is right, it isn't necessary that they last forever. Small PV panels may not be as impressive as a huge solar array on a roof, but their modest milliampere outputs add up to milliampere-hours. A few days of sunshine on such a panel will recharge a battery pack for your hand held transceiver—allowing you to talk with the energy of recycled photons!

From October 1987 QST, p 19:

Alternative Energy—An Overview of Options and Requirements

Part 2: Energy storage is necessary to smooth out natural variations in supply. And what about system safety once your alternative energy plant is up and running?

By Michael Midako, WB6EER
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Some alternative energy sources, such as wind and sunshine, are intermittent and variable in nature. Others may be constant, but of a level too low to meet intermittent peak demands. In all such cases, energy use is determined by the vagaries of nature unless some form of energy storage is employed. One way or another, a means of smoothing out the peaks and filling in the valleys of energy production must be provided.

In hydroelectric systems, this storage may amount to no more than the confinement of water in a reservoir until its energy is needed. Then, opening a valve or sluice gate sets the water in motion, and the kinetic energy in the flow may be tapped by a turbine. Reservoirs work well with water, but are impractical—to say the least—when the energy source is wind or sunshine. A way must be found to store the energy from these sources after it has been converted to electricity.

Hara's solar-powered 2-meter repeater WB6RHR/R, Red Hills (near Shandon), California. Although this installation is not connected to commercial power in real time, it owes its hardware and maintenance to energy-intensive techniques—as do all alternative-energy systems. (photos by WB6EER)

Capacitive Storage

Electrical energy can be stored in capacitors. This is a useful approach when the available charging current is small in relation to a momentary high-current demand, as in photoflash systems, or if the powered system requires voltage at relatively little current, as is the case with short-term memory backup in computer circuitry. Advances in capacitor design allow us to store more and more energy in ever smaller packages, but we are still a long way from seeing capacitors that can compete with storage batteries when the application is one of sustained and regulated discharge.

Electrochemical Storage

Storage batteries provide a practical means for storing large amounts of electrical energy, though it is not really accurate to say that *electricity* is stored in such a battery in a manner akin to capacitive storage. Rather, electrochemically stored energy is invested in a chemical reaction that is reversed when the battery is discharged. The reversibility of this storage reaction is what makes the difference between primary and secondary cells: The electrochemical reaction in primary cells is not easily reversible, disallowing recharging; secondary cells may be discharged and recharged many times.

A wide variety of storage batteries has been developed to meet many storage needs. Sizes and storage capacities range from tiny to enormous. Various battery chemistries are used, depending on the intended service. Which battery you use is determined by the application you have in mind. Size, weight, charge and discharge characteristics, expected lifetime in the proposed service—all of these are important considerations in choosing a storage battery. There is some advantage in using the largest batteries that size, weight, cost and acceptable float-charge load allow: Large batteries mean a large reserve capacity for emergencies or unanticipated use. For a given battery chemistry, life expectancy is generally greater for large batteries than for small ones.

Nickel-Cadmium Batteries

Highly portable low-power applications are commonly powered by nickel-cadmium (NiCd) batteries. These batteries produce a nominal 1.2 V per cell and should survive around 500 charge-discharge cycles. Some NiCd cells can safely sustain rapid recharging, providing an extra measure of flexibility in portable and emergency situations. NiCd cells are produced in the cell packages commonly associated with primary cells (AA, D, C and so on) and can be used interchangeably with primary cells to some extent. It's important to bear in mind, however, that the difference between zinc-carbon and NiCd cell voltages at full charge (0.3 V) makes for significant under-voltage when NiCd cells are series connected to take the place of an equal number of zinc-carbon cells. Perhaps one or two more NiCd cells can be added to such a battery to make up the difference. But the voltage

match is rarely exact in such cases, and addition of more NiCd cells often means substituting too much battery voltage for too little. Since equipment may be damaged by excessive supply voltages, substituting NiCd cells for zinc-carbon units is trickier than it may seem at first—especially if you've added additional cell holders. Let a battery and someone unknowingly installs zinc-carbon cells!

Lead-Acid Batteries

When small battery-powered equipment is used in such a way that the battery is subject to frequent deep discharges, NiCd cells may be the preferred choice. Where deep discharges are only occasional and float-charge current is generally available, a gelled electrolyte lead-acid storage battery should prove more economical in the long run. The nominal cell voltage for lead-acid batteries is 2.0 V.

When it is necessary to power remote sites, especially if they are not vehicle-accessible, 12-V gel batteries rated at about 30 Ah are nearly ideal. Weighing 25 to 30 lbs, they can be transported nearly anywhere with relative ease. Because these are sealed batteries with rugged mechanical characteristics, there is little danger of damage regardless of the contortions that may be necessary to get them to their destination. When higher voltage or greater storage capacity is required, simply use more batteries in series or parallel and distribute the hauling job among carriers or over time. This is infinitely superior to struggling with one giant battery.

Higher power applications, such as operating HF transceivers or household lighting and appliances, require larger batteries. Where the powered site is accessible and power requirements are large, the 30 Ah gel battery is no longer a cost-effective building block. Then, the best compromise between economy and service life is the liquid-electrolyte lead-acid battery.

Automotive batteries are often pressed into this service, more because of their ready availability than suitability for the job. The automotive battery employs a lead-calcium plate chemistry that is satisfactory for brief periods of high-current discharge followed by immediate and complete recharging. Such batteries are not suited to deep-discharge applications where they will be repeatedly drained to a 50% discharged state. In fact, a dozen or so such cycles will reduce the battery's capacity to the point where it should probably not be counted on to start a car. By contrast, batteries designed for deep-cycle service should be good for a few hundred charge-discharge cycles.

This does not mean that automotive batteries are unsuitable for all alternative energy applications. Where the average load current is low and some energy is available to keep the battery float-charged to near capacity most of the time, its useful life may considerably exceed its rating for automotive service. Although the life of such a battery *will* be reduced by deep discharging, the battery will deliver something closer to its rated capacity for the discharge rate in question. Prompt recharging will restore the battery almost to its initial capacity. The self-discharge rate for healthy automotive batteries is lower than that of equivalently rated deep-cycle batteries, so the float-charge current required to keep an idle battery fully charged will be lower for the automotive battery.

Where regular use of higher-power equipment (perhaps 30 W and up) or conversion of battery power to 117 Vac is contemplated, the most practical and economical battery "building block" appears to be the 6-V, 207-Ah units designed for golf carts and similar applications. These are deep-cycle batteries with a lead-antimony plate chemistry. They weigh approximately 70 lbs each and can be moved around fairly easily. For increased storage capacity, they can be connected in series and parallel. Such deep-cycle batteries should have a service life of nearly 10 years if reasonable care is taken in their application.

Large batteries no longer capable or trustworthy in their original service may still do useful work with smaller or less critical loads.

Battery manufacturers consider a battery's useful life to be over when its storage ability has dropped to 50%-80% of its capacity when new. This does not really hold true where the battery has more capacity than necessary for the job. If normal usage of a battery draws only 10% of its rated capacity, it doesn't make much difference whether the battery is 90% as good as new or only 50%. As long as the battery delivers its rated open circuit voltage (no shorted or dead cells) and maintains acceptable voltage under load through the required duty cycle, it is still usefully "alive" for that application. Of course, as a battery ages, its emergency reserve becomes questionable, and overall efficiency is reduced. Eventually, the battery *will* fail; all batteries have a finite life span. The point here is not that we should buy batteries that are much larger than we need, but that large batteries no longer capable or trustworthy in their original service may still do useful work with smaller or less critical loads. This is especially true of older batteries, which can be used to store surplus energy if it is available to trickle charge them.

More Battery Chemistries, Old and New

Earlier this century, much use was made of the nickel-iron chemistry of the Edison cell, particularly because of its lighter weight and tolerance of abuse as compared with the lead-acid batteries of the day. If you can find salvageable Edison batteries, it's quite possible that they can be made to work for you. See the sidebar, "Edison Batteries," for the story.

Looking to where the present blends into the future, research continues in the quest for increased battery life and capacity. Recently, rechargeable lithium cells have made the scene.[2] The dependability of alternative energy systems rests heavily on energy storage, so each improvement in battery and energy management technology is good news for alternative energy planners—especially as the reliability of new technology goes up and costs come down.

Safety in Alternative Energy Systems

As consumers of commercially produced power, we are protected to a considerable degree from electric shock, explosion, mutilation, poisoning and a host of other potential consequences of living in close proximity to the systems and energies that power our civilization. When we take things into our own hands and build energy systems from the ground up, we must consciously build safety in. It is necessary to evaluate hazards and take measures to minimize them.

Next, we'll survey the basic classes of hazards you may encounter in working with the sort of alternative energy techniques outlined so far. This material should not be a substitute for all warnings and instructions that may come with machinery and substances employed in alternative energy work. Nor should it be a substitute for doing personal safety *research*, in the library and face-to-face with experienced people.

The hazards inherent in the production and storage of electrical energy may be divided into three closely related categories: mechanical, chemical and electrical. Some of these hazards are no different from those encountered by any electricity user. Others are more characteristic of complete power systems. As different as they may seem from each other, mechanical, chemical and electrical hazards *are* closely related. A failure or accident in one category is likely to bring about failures in one or both of the others. Such multiple failures can be nearly instantaneous and the consequences can be catastrophic.

[2]Nathan Bowker and Christopher Ballard, "The Magic of MOLI," Jun 1987 *QST*, pp 22-25.

Edison Batteries

First marketed in the early 1900s, the nickel-iron alkaline Edison cell has accumulated a reputation for capacity and indestructibility that is only partially justified. It is *not* the perfect storage cell, but it does have some interesting qualities. Batteries of Edison cells were designed to survive rough mechanical abuse in railroad lighting and vehicle propulsion service. Largely because of the strong, lightweight construction of its steel case and its rugged internal structure, the Edison battery achieved this objective with a better power-to-weight ratio than could be attained readily by the lead-acid batteries of the time.

The construction and chemistry of the nickel-iron cell is such that it can survive abuse that would be fatal to a lead-acid cell. As long as it is not drastically overheated, the Edison cell can be overcharged to the point of vaporizing all of the electrolyte and no great harm will result. Nor will the cell be harmed by being left in a totally discharged condition. I know of used Edison cells that recovered a good percentage of their original capacity upon being filled with distilled water and run through a few charge-discharge cycles—after having been dry and totally neglected for over *40 years*.

Now for the bad news. As compared to lead-acid cells, the Edison cell has a high internal resistance and a high self-discharge rate. Thus, voltage regulation during load variation is poor, and the cell shows a continuous loss of voltage throughout its discharge cycle—from nearly 1.4 V at full charge to 1.0 V at the bottom of the cycle. Hydrogen and oxygen are vented continually, though to varying degrees.

Edison cells employ a potassium hydroxide electrolyte. This is a strong base and must be handled with caution. Acids and acid-contaminated tools should *never* be used in or around Edison batteries—something to keep in mind if your battery "stable" is to include both lead-acid and Edison cells.

A hydrometer is not of much use in determining the state of an Edison cell because the specific gravity of the electrolyte changes little between the charged and discharged states. Cell voltage, charging time and charging current are the best indicators of charge for Edison batteries.

Terminal voltage in the discharged condition for a single Edison cell is considered to be 1 V. New Edison batteries had an expected lifetime of 2000 charge-discharge cycles. Most of these batteries were probably used by railroads for passenger car lighting and trackside signaling, although many saw service in domestic wind power installations. Despite their age, however, Edison batteries may still be found. Many of their cells will undoubtedly be in salvageable condition (see photo).

A battery of six Edison cells as collected on a salvage expedition. Most of the cells are good, but the original wooden rack is in bad shape. These are A-8 cells, originally rated at around 220 Ah each.

If you come across an odd-looking battery like that shown in the photo, don't assume that it is dead and gone. If the steel case of a given cell is intact and the poles are not internally shorted or shorted to the case, it is quite possible that the cell can be revived. Cases of adjacent cells in an Edison battery must be insulated from each other or electrolytic action will eat through them in short order. (In an Edison cell, the steel case is isolated from both poles but common to the electrolyte.) A socket wrench and a good gear puller are essential for disassembly of Edison batteries. Details on the care and feeding of Edison batteries can be found in older electrical engineering handbooks.[1]

[1]Greatly detailed information on Edison and other secondary cells may be found in George Wood Vinal, *Storage Batteries*, 2nd ed. (New York: John Wiley and Sons, 1930).

The sidebar, "Harmless," offers an example of the kind of nasty multiple failure that can happen around an alternative energy installation. Although the chain of events depicted there may seem farfetched, it isn't. When you achieve long periods of accident-free alternative energy production, you *won't* have wasted your time anticipating and guarding against the worst!

Mechanical Hazards

Moving parts, especially gears, vee belts, pulleys, wind turbine propellers and the like, should all be made inaccessible to accidental contact. This is usually accomplished with covers and enclosures. When such moving parts must be exposed, they should be located out of reach. A wind turbine should not be able to touch anyone on the ground or working on its tower.

Towers should be designed and supported to withstand worst-case weather conditions for the area. They should receive regular inspections and maintenance as needed. When in doubt, consult a structural engineer. Towers are attractive nuisances, so they should not be climbable by children or passersby.

Persons developing any energy resource must take a certain responsibility for their safety and that of their neighbors.

Chemical Hazards

All motor fuels and their vapors are flammable and potentially explosive. They must be handled in suitable containers, lines and fittings. Most fuel vapors have distinctive odors, so use your nose! Don't ignore what your sense of smell tells you. Track down and repair leaks. Never store fuels near operating engines or sources of open flame and sparks.

Internal combustion engines produce carbon monoxide gas as an exhaust product. This is a colorless, odorless and lethal substance. Do not breathe exhaust fumes; also, do not risk operating engines in enclosed spaces unless exhaust fumes are properly vented through a gas-tight system. Even with a good exhaust system, it's good

Harmless

An industrious mouse enters the battery compartment of an alternative energy system. Shuffle, sniff. No loose scraps worth taking—just a foot-long piece of bare no. 10 wire carelessly abandoned in the framing of the compartment two years ago. Exiting the compartment, the rodent shoulders the scrap aside, causing it to fall across the terminals of a 12-V storage battery. There is an immediate electrical failure as the wire welds to the battery terminals, shorting the system. The wire reaches red heat in a matter of seconds. As it glows brighter and begins to melt, the wire slumps onto the plastic battery case. The case melts like butter under a hot knife.

At this point, the electrical failure is over. The wire melts through the battery case with a sizzling arc that causes the hydrogen and oxygen within the battery to unite with explosive force. The explosion rips the already damaged battery open, spewing sulfuric acid, acid vapors and hot metal all over the battery compartment.

With luck, the problem ends here, with no fire climbing the walls and no injuries—just a terrible mess to clean up. But don't count on it. A chance encounter with a harmless scrap of wire and a mouse has already blown up your battery. Why should chance stop there?

Such a series of events may seem highly improbable. But trusting to probability implies taking chances—in other words, playing odds. And that's exactly what *not* to do when building safety into an alternative energy system. Dangerous system failures are possible unless care is taken to make them *impossible*. You must build safety in.

insurance to keep a carbon monoxide alarm in the engine room.

Engine exhaust systems can emit burning gases and hot carbon particles, both of which can ignite dry materials in the vicinity of the exhaust outlet. When internal-combustion-engine driven generators are to be used outdoors under dry conditions, use spark arresting mufflers or spark arresters approved by the US Forest Service. Clear a ten-foot radius of bare dirt around the generator and *keep* it clear. Have a shovel and fire extinguisher nearby and in plain sight.

Whether they're acidic or alkaline, battery electrolytes are nasty substances. They can corrode metal, creating both mechanical and electrical problems. They can destroy clothing in short order, and their activity does not stop when they get to the flesh underneath. Soft tissues such as eyes, are particularly prone to rapid damage from exposure to battery electrolytes, so wear eye protection when working around batteries. Keep some means of flushing away accidental exposures at hand; a garden hose will do. Don't wear your best clothing when working with batteries—some exposure to electrolyte is almost inevitable. The evidence may not appear until that special shirt comes out of the washer looking like cheesecloth!

Avoid panic by having emergency procedures well in mind. Your flesh won't dissolve right off your bones if you *do* get electrolyte on it, so don't go into shock. Just start flushing the affected area immediately. If garments are saturated, get out of them.

Storage batteries (except for completely sealed recombining types) emit hydrogen and oxygen gases, particularly under heavy charging and overcharging. This is a highly flammable, explosive mixture. Although hydrogen is much lighter than air and tends to dissipate rapidly, it cannot do this in confined spaces—such as the space between the electrolyte surface and the filler cap of a battery. Dangerous concentrations of hydrogen can accumulate here. Thus, checking the electrolyte level by match light or "testing" a battery by drawing sparks across its terminals are dangerous techniques and should *never* be used.

Dangerous system failures are possible unless care is taken to make them impossible. You must build safety in.

Storage batteries also tend to vent corrosive vapors that can damage delicate electronic equipment. If vented batteries are used indoors, the vents should be extended to the outdoors with plastic tubing. The best practice is to provide storage batteries with their own well-ventilated compartment or room.

Electrical Hazards

Electric shock is to be avoided at all costs. Shock danger from 12-V dc systems is minimal, but as system voltage approaches 32 V, it's possible to get "bitten" and even be electrocuted if conditions are just right (or wrong!). Both storage batteries and solar panels connected in series can add up to shock potential in short order. Remember that the output voltage from solar panels is much higher with no load than it is when a load is connected. Where sinusoidal ac energy is concerned, thinking in terms of RMS voltage can be deceptive, because ac peak voltage works to overcome your skin resistance—and *peak* voltage in a sine wave exceeds RMS by a factor of 1.414.

Current Kills—But It Also Burns

Even small storage batteries can deliver high currents sufficient to bring small conductors to red heat, creating potential for fire and burns. Larger batteries, such as those found in automobiles and alternative energy storage systems, can deliver hundreds of amperes. Such currents can heat and melt large conductors. Rings, bracelets and wristwatches should never be worn by people working with electrical systems for this reason. Electrocution may be the first danger that comes to mind when considering the wearing of metal jewelry, and it should never be ruled out, of course. But stories of fingers amputated and cauterized by a white hot ring welded across a high current source are not fables—it can happen to *you*.

Protect battery terminals from short-circuits. Exercise extreme caution if you must work around batteries with metal tools. Always keep one terminal covered to avoid the possibility of a short circuit.

Modern battery cases melt readily even at soldering temperatures (360-460° Fahrenheit for common solders). These cases also deteriorate rapidly in sunlight,

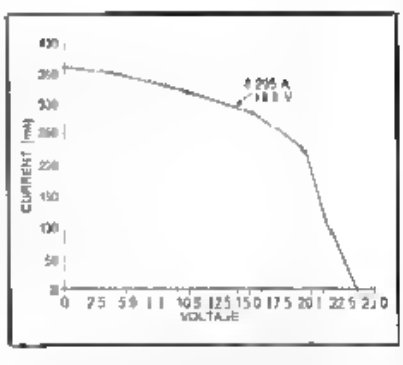

Fig 2—Voltage across the terminals of a "12 V" solar panel varies considerably with load, and this must be allowed for in the design of a solar energy system. (The graph shows voltage versus current for the 5-W Sovonics panel described in the article called out at Note 1 in Part 1 of this article.)

leading to embrittlement and cracking. Keep them out of the sun and handle them with care.

Fusing and Load Switching

Fuses are essential insurance for electrical safety. Fuses or circuit breakers rated to handle full load current should be placed as closely as possible to the battery. Great care must be taken with insulation and dress of the wiring from battery terminals to fuses. Since high currents at low voltages are involved, low-resistance connections to fuses and breakers must be provided. Further fusing of subsystems as appropriate to their individual current demands can be installed at a convenient location farther from the battery.

In switching and fusing a photovoltaic system, bear in mind that "12-V" solar panels may produce more than 20 V across an open circuit or high-resistance load (see Fig 2). This could have disastrous consequences for equipment should the line from the PV array to the battery open with equipment still connected to the PV array. If at all possible, meters should be used to monitor charging current, load current and battery voltage in an alternative energy system. Then, proper operation of the system can be confirmed at a glance.

Conclusion

If you find yourself inspired to become involved with alternative energy projects, you'll discover a wealth of literature devoted both to specific and general topics in the field. The few references I've listed in the bibliography will help get you started. It's also quite likely that you can share ideas and questions with someone in your own area who is working commercially or privately with some aspect of alternative energy. Such people may well be the most valuable untapped resource you'll find as you work to develop an operational energy system.

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From July 1989 QST, p 31.

Operate Your Station With Power from the Sun!

Here's a report on this *hot* technology, with the information you'll need to design your own solar-powered station.

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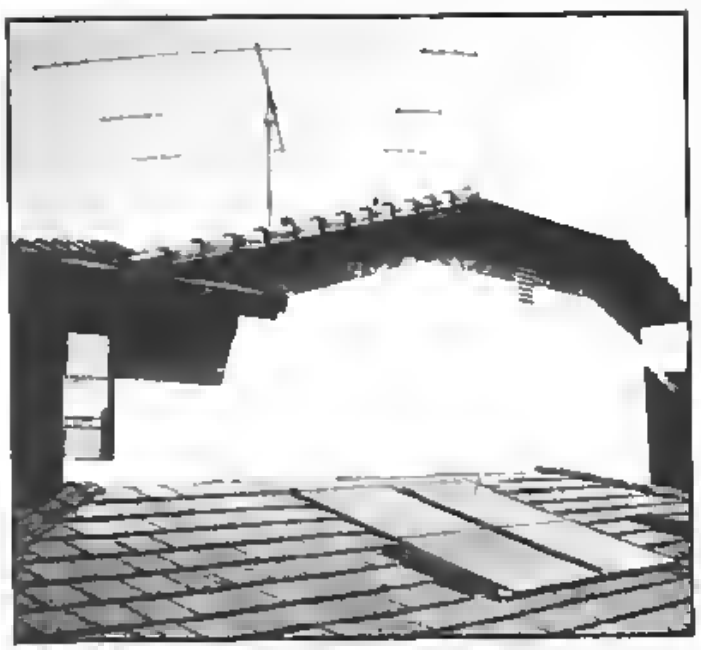
I'm not a dyed-in-the-wool ham radio operator when it comes to RF—I'm more of a tinkerer in electronics. Although I have held a license since 1954, I have, until recently, derived more satisfaction from designing receivers and transmitters—and getting them to work—than from operating them on the air.

With the sunspot cycle starting upward again, I finally broke down, put savings in pocket, drove to the ham store and purchased a new do-operated transceiver. Because I did not have a sufficiently large (20-A) power supply for this radio, and because I was in a hurry to try out the new rig, I borrowed the storage battery from my motor home. This battery has a 55 ampere-hour (Ah) capacity, can be deep discharged, and appeared to be plenty large for this application. It has operated the rig for over a year without trouble.



Solar panels (lower right) and a tribander decorate the rooftops at KG6JA.

I recently read up on developments in alternative energy sources such as wind, motion (water) and sun that supply energy to power an entire household. It occurred to me that the use of solar energy to charge a battery capable of powering my ham station would be an interesting and affordable experiment. Much to my surprise, I found that technology in the manufacture of solar electric cells has moved far enough forward that you don't have to live in space, in Florida or in California to benefit from solar energy! Solar-cell efficiency is such that solar cells can provide sufficient energy to be usable in areas of the country where sunshine is less abundant than in the Sunbelt. In fact, these cells even provide electricity on cloudy days.[1]

Capturing Energy from the Sun

The electric effects of light on certain materials have been known since long before the invention of the transistor. Materials such as cadmium sulfide and selenium exhibit altered electrical behavior when they are exposed to light. Early in the development of transistors, it was discovered that transistors not encapsulated in lightproof housings were sensitive to light. The reason for this is that photons striking a base-emitter junction cause the movement of electron-hole pairs in the junction—just as injecting a forward base-emitter current does. An increase in collector current is the result. This discovery was later put to good use in the development of the translation of light energy to electromotive force, *photovoltaic conversion*, often called PV or called *PV conversion*.

A solar cell is a very simple semiconductor. Solar cells are, in fact, large-area semiconductor diodes. A cross-section of a solar cell is shown in Fig 1. Simply explained, when the photons contained in light rays bombard the barrier of this semiconductor, hole-electron pairs inside this P-N junction are freed, resulting in a forward bias of the junction, just as in phototransistors. This forward-biased junction can deliver current into a load. Because the exposed area of a solar cell can be quite large, the forward current produced can be substantial. It follows that the output current of a photocell is directly proportional to the rate of photon bombardment, and thus to the exposed area of the photocell.

Types of Solar Cells

Originally, solar cells were made by cutting slices of grown silicon-crystal rod and subjecting them to doping and metallization processes. These solar cells are called *monocrystalline* cells because each unit consists of only one crystal plate. The shape of these cells is the same as that of the silicon rod from which they are cut: round. A slice of this material with an area of 2 inches can be made into one photocell, but a slice of this size could also be used to

[1]Notes appear at end of article.

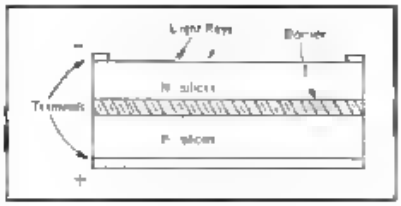


Fig 1—Cross section of a PV solar cell

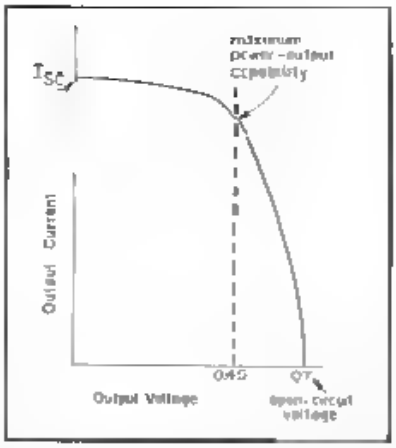


Fig 2—This load curve for a PV solar cell shows that maximum power delivery from a solar cell occurs at approximately 0.45 V output. I_{SC} is the short-circuit current.

produce upwards of a thousand transistors! The cost of these early solar cells was way beyond the means of common folk, and could only be justified for use in space research and other highly critical applications.

Techniques for the manufacture of two other types of PV cells have been developed since monocrystalline cells were first produced. The newer cells are *polycrystalline cells* and *amorphous cells*. Polycrystalline cells are typically manufactured as rectangular blocks of seemingly randomly arranged silicon crystals from which the cell plates are cut. These cells can be recognized by their shape and their random pattern and colorful surface. Polycrystalline cells are less expensive to manufacture than monocrystalline cells.

In the mid 1970s, researchers began to experiment with the manufacture of PV cells by depositing a thin film of doped silicon on an economical but stable substrate, such as glass. In 1975, these efforts paid off, and today the result, amorphous cells, are used in the production of calculators, watches, security systems, automatic gate openers, electric fences, wireless freeway telephones, battery chargers in automobiles and recreational vehicles, and, of course, in ham radio. At first, amorphous solar cells were not very efficient and exhibited rapid degradation with time. Most of these problems have since been solved, and reliable amorphous PV panels are available from many manufacturers. These panels come in several forms, mounted on thin glass, framed, and even mounted on flexible substrates such as steel.

Amorphous cells are relatively inexpensive to manufacture. They do not, however, spell the end of crystalline cells. Crystalline cells still offer the highest efficiency. The best shopping comparison you can make when purchasing PV solar panels is to compare power output per dollar, and then select a vendor who offers a good warranty and good customer service.

Solar-Cell Specifications

Depending on construction, each cell has an open-circuit output, when exposed to the sun, of 0.5 to 0.8 V. One may have expected this, because each cell is the electrical equivalent of a forward-biased silicon diode. The output voltage drops somewhat when current is drawn from a solar cell. Fig 2 shows the typical voltage v current relationship of a solar cell. This is called the cell's *load curve*. Open-circuit voltage is approximately 0.7, and output voltage at optimum load is nominally 0.45. Output current is maximum with shorted output terminals. This maximum current is called the short-circuit current, or I_{SC}, and is dependent on the cell type and size. Because a cell's output current remains relatively constant under varying load conditions, it can be considered to be a constant-current source. The point on the load curve where maximum power can be drawn from the cell is indicated in Fig 2.

Just like batteries, solar cells may be operated in series to increase output voltage, and/or in parallel to increase output-current capability. Several manufacturers supply arrays or panels with a number of cells in a series-parallel hookup to be used, for example, for battery charging.

Techniques have been developed for the construction of amorphous cells whereby the cells are manufactured in series by cutting metal layers that have been vapor deposited on the amorphous silicon mass. This cutting is done with a laser. Cell width in such panels may be up to several feet, and the output-current capability of these relatively economical panels is excellent.

PV-cell efficiency varies. Monocrystalline cells have efficiencies up to 15%; polycrystalline cells, 10 to 12%; amorphous cells, 6.5 to over 10%, depending on the manufacturing process.[1]

The output power of solar arrays or panels is specified in watts. Typically, the listed wattage is measured at full exposure to sunlight, at a nominal potential of 7 V for a 6-V system, 14 V for a 12-V system, and so on. You can calculate the maximum current that can be expected from a PV panel by dividing the specified output power by the panel voltage.

The cost of solar panels has decreased significantly in recent years. Basically, you can expect to pay anywhere from about $8 to $15 per watt, depending on quantity, size, construction and efficiency of the panel.

Storing Solar Energy

Because the sun doesn't shine 24 hours per day at any location in the US, some means of storing collected energy must be used. Batteries are commonly used for this purpose. Battery capacity is generally expressed in ampere hours (Ah) or milliampere hours (mAh). This rating is simply the product of discharge current and discharge time in hours. For example, a fully charged 500-mAh NiCd battery of good quality can deliver a discharge current of 100 mA for a period of 5 hours, or 200 mA for 2½ hours, before recharging is required.[2] Three types of rechargeable batteries are commonly used:

- Nickel-cadmium (NiCd) batteries: NiCds are mostly used in relatively low-energy applications such as hand-held transceivers, scanners, etc. The development of consumer electronics has contributed to the rapidly increasing availability (and somewhat-less-rapidly decreasing cost) of NiCds. Major advantages of NiCds: They are hermetically sealed, operate in any position and have a good service life (several hundred charge/discharge cycles), if they are properly maintained.
- Gelled-electrolyte lead-acid batteries: These hermetically sealed batteries are available in capacities from below 1 Ah to more than 30 Ah. They are ideal for supplying energy to a ham radio station, but their cost (for capacities above 10 Ah) is rather high. For portable and QRP stations, though, this type of battery is difficult to beat. The cells can be operated in any position, but should be charged in an upright position. If properly maintained (no deep discharges—cell-polarity reversal is possible under these conditions—and they are stored in a fully charged state), gel cells last a long time (500 or 10 cycles). I operate a small 10-W portable CW station from a 12-V, 6.5-Ah gel battery with good success.
- Ordinary lead-acid batteries: These are available in the standard automotive version, in the marine/RV deep-discharge versions and in the golf-cart variety. Differences: Automotive batteries usually fail following several deep-discharge cycles (because of the thin plate and insulation materials used in their construction), resulting in premature internal short circuits. Golf-cart and marine/RV batteries have thicker plates with more rigid insulation between them, so these batteries can withstand deeper discharges without plate deformation and internal failure. Deep-discharge batteries provide the best value in a ham station. Some of these batteries require attention (the electrolyte level must

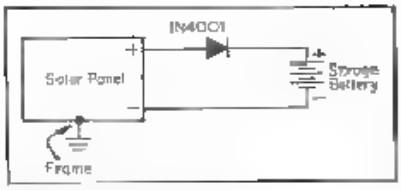


Fig 3—Basic solar-panel-to-battery connection. The series diode in the panel's output line keeps the battery from discharging into the panel.

be maintained), and they last longest when kept charged. Because these batteries use a wet electrolyte (water), and most of them are not hermetically sealed, they must be kept upright.

A Typical Application

Here's a practical example of how to calculate power requirements for a PV-powered ham radio station. The first thing to do is define the power demand. Assume that you use a 100-W rig. (We'll also assume that 100 W is the peak power consumption, and occurs only during CW operation and SSB voice peaks when a 13.6-V nominal supply [a fully charged battery] is provided.)

The most reliable way to calculate realistic power requirements is to determine the power used over a longer period of time—say, a week or a month. Because most of us have more or less recurring weekly habits, we'll take one week as the base period. (You can substitute your own numbers to adapt these calculations for your rig, under your operating circumstances.) Assume that the rig is turned on five days of the week for two hours on each of these five days. Of each two-hour period, 1½ hours is spent listening, and transmitting takes the remaining half hour. Assume that the current consumption of the transceiver during receive is 2 A; during the 100-W peaks on transmit, current drawn is 20 A. The owner's manual for your rig should give the maximum dc current drain. The average current consumption during SSB transmitting is only about 4 A. Therefore, we need a battery that can supply a peak current of at least 20 A and an average current of 4 A. Now calculate the total energy consumed in ampere hours over a one week period.

Receiving: 2 A × 2½ hours/day × 5 days = 25 Ah.

Transmitting: 4 A × ½ hours/day × 5 days = 10 Ah.

The total energy used per week is 25 + 10 = 35 Ah, or per day (average) is 35 ÷ 7 = 5 Ah. If we had a perfect system, all we would need to do is supply 35 Ah per week (5 Ah per day) to the battery. In practice, imperfections in battery construction cause some loss (self discharge), for which the charging system must compensate, as you'll see.

Next, calculate the minimum battery capacity required for this application. The system should be designed so that sufficient energy is available to run the equipment for two consecutive sunless days (this is rather arbitrary—some locations are worse than others in this regard). Because these sunless days could be days on which you want to operate, and because it's not a good idea to discharge a battery to less than 50% of its capacity (for maximum battery life), this battery must have a capacity of at least 2 (days) × 5(Ah) ÷ 0.5 (for the 50% charge capacity left after 2 days without sunshine) = 20 Ah. If your location is likely to be without sunshine for as much as an entire week, the battery requirement is 7 × 5 ÷ 0.5 = 70 Ah. Add about 10% to this number to compensate for self discharge and other losses. (Typically, this means you'll buy the next larger-size battery than the initial calculations indicated.)

What does it take to keep this battery sufficiently charged? Here again, some rules of thumb help in the calculations.

First, estimate the average number of hours of sunshine per year in your area. This information can be found in an almanac. As a guide, average annual sun exposure is approximately 3200 hours per year in the Sunbelt, less elsewhere (down to about 1920 hours per year in the far northern parts of the US).

Your PV solar panel will most likely be mounted in a fixed position, but should be at an optimum angle with respect to the earth. This varies from about 30° in the summer up to about 60° in the dead of winter. Fixed-mounted solar panels cannot pick up maximum energy from the sun, for obvious reasons. Of course, you could build some kind of solar-tracking mechanism to circumvent this obstacle, but that's beyond the scope of this article (and beyond the ambition of most people I know). If you need to collect more solar energy, it is much easier to simply add another solar panel! In practice, you can only count on panel exposure for about 70% of the total sunlit time, which is anywhere between 1340 and 2240 hours per year (between 26 and 43 hours per week), depending on where you live.

The remaining system planning is easy. Our earlier calculations showed that the solar cells must replenish 35 Ah per week, plus 10% to compensate for losses, or about 38.5 Ah of battery capacity. With solar *energy* available in the Sunbelt for 43 hours per week, the required charge current is 38.5 Ah ÷ 43 hours of sunshine = 0.9 A. In the northern part of the US, this is 38.5 Ah ÷ 25.3 hours = 1.5 A. Now, find a PV panel that can deliver this current under load.

In the 12-V system described here, the PV panel operates, with a fully charged battery, at about 13.6 V, plus the voltage drop of a series diode. The basic hookup is shown in Fig 3. With a fully loaded panel voltage of 14, a panel rated at 21 W (14 V × 1.5 A) is required in northern climes. In practice, this power can be obtained from good-quality solar panels with a surface area as small as 5 square feet. If you live in the Sunbelt, you need only 12.6 W (14 V × 0.9 A) of PV energy.

Using this basic method, you can calculate the electrical and mechanical dimensions of almost any solar installation, just substitute your power needs into the equations shown here. The sidebar, "Calculating Solar-Cell and Storage-Battery Needs," conveniently shows the required calculations in tabular form.

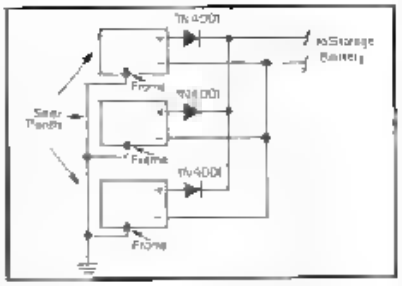


Fig 4—Connecting panels in parallel is a good way to increase output current from a solar panel array. A diode is connected in series with each panel to protect the panels from the battery and from the other panels.

Some Practical Hints

PV panels can be wired in series to provide increased output voltage. If the total output of the cell array exceeds 20 V, wire shunt diodes across each PV cell. Similarly, PV panels can be wired in parallel to yield increased output-current capability. In this case, use a series diode for each panel, as shown in Fig 4.

When hooking up PV panels to a storage battery, always use a series diode to prevent discharge of the battery into the panels. A Schottky diode can be used in applications where it is important to maintain the lowest voltage drop (and minimum loss) of charge current.

If you live in an area where battery overcharging may occur, take precautions to prevent battery overcharging and related gas discharge inside the battery. Several manufacturers supply simple charge regulators that serve this purpose by disconnecting the PV panel from the battery when the battery is fully charged. Some of these chargers allow charging to resume when the battery has reached a measurable level of discharge. Note: These values are only valid for lead-acid batteries; an entirely different set of charge criteria exists for NiCds.[4]

Installing Solar Panels

If you plan to permanently install PV

Calculating Solar-Cell and Storage-Battery Needs

Calculation of PV solar-cell and storage-battery requirements is easy using this form. See the text for additional information.

Solar Cell

1) Current drain during receive † _____ A
2) Number of hours of receiver operation per week. _____ h
3) Multiply (1) × (2) _____ Ah
4) Measure peak current drain during transmit.† _____ A
5) For SSB, enter 0.2, for CW, 0.5, else, 1.0 _____
6) Number of hours of transmitter operation per week _____ h
7) Multiply (4) × (5) × (6)* _____ Ah
8) Add (3) + (7): _____ Ah
9) Number of hours of sunshine per week (see text)* _____ h
10) Required solar-panel current [(8) ÷ (9)]* _____ A

The solar panel you select should have at least as much current capability as shown in line 10.

Storage Battery

11) Maximum number of days of operation without sunlight: _____
12) Daily power requirement [(8) ÷ 7] : _____ Ah
13) Stored-energy requirement [(11) × (12)] _____ Ah
14) Maximum stored-energy requirement [(13) × 2.2]: _____ Ah

The battery you select should have a capacity at least as large as the total storage requirement found in (14), or the numerical current capacity of the battery (in Ah) should be at least twice as large as the numerical value found in (4), whichever is larger.

†These values are usually given in the transceiver instruction manual.

panels, consider mounting them at ground level on a simple wooden or metal frame, or mounting them on the roof. Roof mounting is more appropriate if you have a roof that slopes at the correct angle (30 to 60°—see the title photo), and in the right direction (anywhere between a little east of south and southwest is acceptable). The easiest way to mount panels permanently is with a silicone adhesive, such as RTV. First, mount series diodes on the back of each panel. Attach color-coded wires to the panels' negative terminal or wire, and to the cathodes of the diodes. If you're using more than one panel, wire them in parallel so that you only need to run one set of leads from the panels to the battery. Secure the wires and diodes with small strips of tape and apply a blob of silicone adhesive to each diode and solder joint.

If the solar panels are going to be located in an area where they might be subjected to lightning, it is especially important to ground the metal frames of the solar panels. Use separate wire for this ground—*do not* combine the panel-frame ground with one of the power leads!

After you have determined where the panels will be positioned, lay them upside down and squeeze a bead of silicone adhesive onto the back of each panel frame. Turn each panel over, lay it on the roof, and tape it down until the adhesive has dried. If you want to go the extra mile, you can mount the panels on blocks so that air can circulate under the panel. PV solar panels when cool, have slightly higher output than hot panels.

Methods of securing solar panels to wooden or metal frames vary with frame designs; panels may lay in the frame, or you may elect to use brackets and/or bolts. The advantage of a frame mount (as opposed to a more permanent mounting scheme, such as adhesive), is that you can adjust the angle of the panels with respect to the ground, so that you can align the panels for maximum efficiency.

Notes

[1] I use *provide* rather than *generate* to describe the process by which solar energy is transformed into electrical energy. Of course, the energy is generated by the sun; solar cells convert photon energy into electrical energy, and thus *provide* electricity.

[2] For reference, maximum theoretical PV-energy-transfer efficiency is in the 22 to 25% bracket. The optimum has not been reached, but it is pretty well approached (to within an order of magnitude).

[3] Ampere-hour battery ratings are not absolute. You generally cannot, for instance, use a battery with a rating of 2 Ah to supply 10 A for 0.2 h (2 Ah ÷ 10 A), or even 4 A for ½ h. Materials used in battery construction are not made to withstand the heat generated during such extreme operation.—*Ed*

[4] I intend to describe a simple charge regulator for lead-acid batteries in a future article. Drop me an SASE for more information about charge regulators.

From November 1987 QST, p 41:

Free NiCd Cells

☐ Many cities and towns have an electric razor repair shop that replaces NiCd batteries in rechargeable electric razors. The razors I've seen contain two NiCd cells. Often, only one of these cells has failed, but both cells are replaced when repair time comes. I talked a razor repair person out of a box of such rejects and got 40 NiCd cells. Twenty of these charged perfectly on the first try! Brief application of heavy overcharging current to the rest of them netted another dozen usable cells. Free NiCd cells? Check your friendly electric razor repair shop.—*Bob Baird, W7CSD, 3740 Summers Ln, Klamath Falls, OR 97603*

From March 1988 QST, p 41:

A Deep-Cycle Battery as an Emergency Power Source

☐ After I acquired a size 27, deep-cycle lead-acid battery as an emergency power source for my 2-meter transceiver, hams on the local repeater advised me on how to keep the battery charged. "Connect a variable dc supply in parallel with the battery and set its output voltage to 13.6," they said.

The current capability of my power supply is insufficient for such service. The supply can source the 4 to 6 A required by the rig during high-power transmit, but is rated at only 3 A for continuous duty. Connecting the supply directly in parallel with the battery *and* the transceiver would, at times, result in current drain exceeding the supply's continuous-duty rating.

Fig 2 shows my solution to this problem. Charging current with this circuit is 1 A or less, and the supply can still power the transceiver. Installation of a jumper across points A and B applies the full battery voltage to the transceiver if this is needed during an extended power failure. —*George Hopkins, KF6GL, 521 S Castlerock Ter, Sunnyvale, CA 94087*

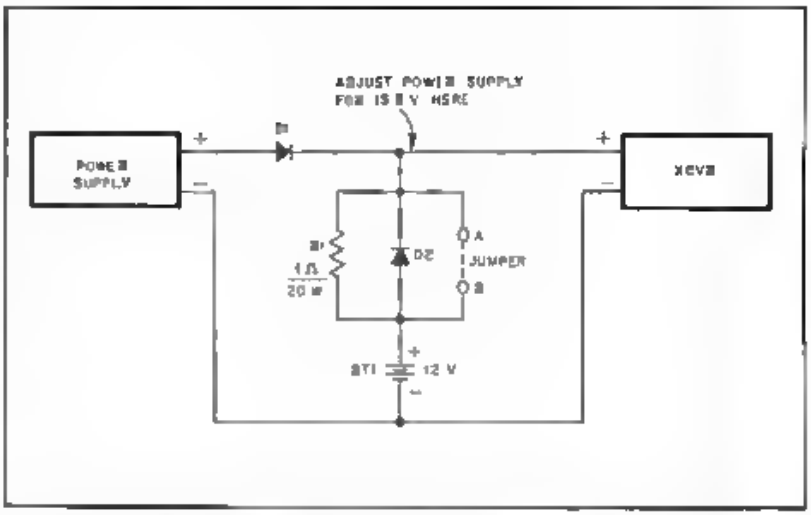


Fig 2—KF6GL's power supply/charging circuit. In this application D1 and D2 are 6-A, 600 PIV diodes (Motorola MR756, ECG5815 or equivalent). R1 is a wirewound unit and BT1 is a size 27, deep-cycle, lead-acid storage battery. The jumper is used only during power failures (see text). The power supply is normally turned on only while the station is attended. If tailoring this circuit to your application, use conservatively rated components.

From June 1988 QST, p 47:

A One-Shot Timer for Battery Charging

☐ One of the problems associated with rechargeable batteries is that of charging duration. This is particularly evident when the charging period is longer than the interval between "home from work" and "back to work"! In such cases, another member of the family must remember to unplug the charger at the appropriate time. My solution to this problem is a one-shot timer. Here's how to build such a timer for around $10.

Two parts are required: a 120-V neon lamp assembly and a motor-driven lamp timer capable of a timing interval at least as long as the charging period required by your battery. (Both are available in several forms from Radio Shack®.) The motor in such timers usually actuates a switch that breaks the hot side of the ac line for appliance control. Modify the timer as shown in Fig 1. Open the timer and locate the motor lead connected to the hot side of the ac line (point A in Fig 1). Move this lead to the appliance-socket side of the timer switch, S1 (point B in Fig 1). Next, mount the neon lamp assembly at any convenient place on the timer housing, and connect it between the hot and neutral terminals on the timer's appliance outlet. Reassemble the timer.

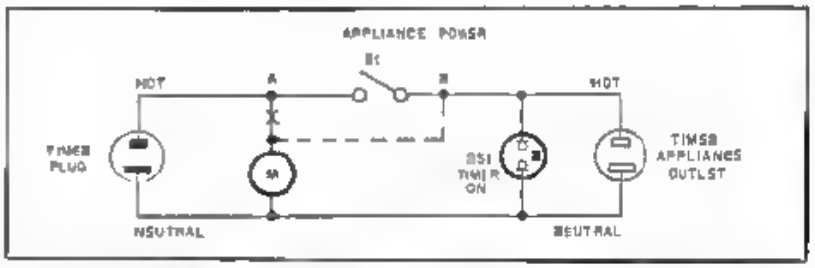


Fig 1—Dennis Cripps' modified timer turns itself and its associated battery charger off at the end of the charging period. Modification of the timer consists of moving the hot motor lead from point A to point B, and installation of a 120-V neon lamp, DS1, between hot and neutral on the timer's appliance outlet.

Connect your battery charger to the timer. Set the timer's on and off actuators to turn on the charger for the charging time required by your battery. To turn on the timer motor and your battery charger, rotate the timer dial until DS1 lights. When the set time has elapsed, your charger *and* the timer motor switch off. Result: a charged battery that won't be overcharged if forgotten.—*Dennis Cripps, N3FIW, 318 N Dillwyn Rd, Newark, DE 19711*

From May 1976 QST, p 25:

Power Amplifier Development with Your Transisitors

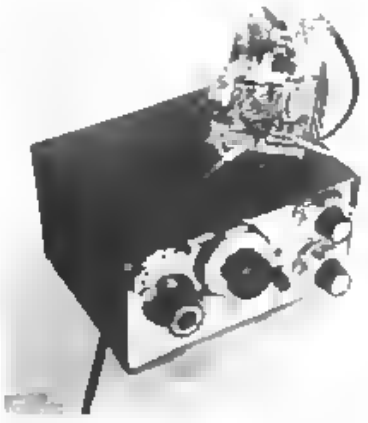
Simple test equipment and methods for making-do with devices on hand, on frequencies you want to use.

By Adrian Weiss, W0RSP
833 Duke St 83
Vermillion, SD 57069

One of the more exciting phases of ham radio today is the use of rf power transistors in transmitter amplifier stages. Solid-state design has obvious weight and power-drain advantages, especially in gear that may be used for mobile or portable operation. Development of balanced-emitter rf power transistors, virtually burnout proof and superior to earlier types in regard to stability, gave great impetus to use of all-solid state equipment in both the hf and vhf ranges.

For the amateur who wants to do other than make exact copies of described equipment, a problem has been lack of understandable information that will permit him to work out transmitter designs for transistors he may have on hand or be able to pick up at moderate prices. Even when good information is available, it may be for only the vhf range, or the circuits described may not necessarily be the best available for amateur-band use. Unlike vacuum tubes, solid-state devices may exhibit wide variation between individual units of the same type. This is in part the result of applications design for top-quality production runs intended for military or space use, whereas the amateur may have to contend with second- or third-level quality. There is also the matter of the practical unreliability of mathematical calculations used in solid-state amplifier design. Johnson and Artigo have noted that competent engineering can produce "ball-park" errors ranging from -22 to +25 percent between calculated values and those that actually work.[1]

Solid-state 40 meter amplifier, built by the author for use with his QRP rig, was tested and adjusted using the methods he describes



Assumptions

The objective here is to allow the average amateur to circumvent the above obstacles, by placing emphasis on the actual device on hand through in-circuit measurements made during amplifier development. The method is based on several general assumptions which will hold in most cases. A reader unfamiliar with solid-state amplifier basics is encouraged to study papers by Hanson, Hayward, Hejhall, and others.[2]

It is assumed that the base input impedance of the amplifier will be quite low, in the range of 1 to 15 ohms. The input matching network must be able to transform this low impedance to whatever is present at the output of the driver stage. This could be 50 ohms, as in using an amplifier with a separate exciter such as one described by the author in an earlier article,[3] and shown in the photograph, or some higher value if the exciter is to be an integral part of a complete transmitter. A reactive component will be present in the base input impedance, so the interstage matching network must tune the base input circuit to resonance, as well. The amplifier will operate properly only when both conditions are satisfied.

Any balanced-emitter device will have an absolute minimum gain of about 6 dB if operating properly. Efficiency will be 45 to 65 percent. At least 8-dB gain is expected normally. On this basis the drive required for 10-watts output is 1.25 watts. In practice, the writer has found the 2N5590 can be driven to about 12.5-watts output with 1 watt of drive. In another application the 2N5590 delivered 5.5 watts of clean output with only 220 mW of drive – about 14-dB gain. A word of caution is in order here. Maximum efficiency is obtainable only at the collector voltage specified by the manufacturer. Don't expect high efficiency if a 28-volt device is operated at 12 volts.

Practical Circuit Details

Hayward discussed choosing values for the base swamping resistor, collector rf choke, bypass capacitors, and other components of the typical Class-C amplifier. Bearing in mind that these circuits are not official "dogma," the reader is advised to familiarize himself with them. There are several usable circuits, descriptions of which can be found in the references and in the RCA *RF Power Transistor Manual*. The author prefers the input network shown in Fig. 1, because it will yield practical component values in nearly all cases.

[1] This and all subsequent footnotes will appear at the end of this article.

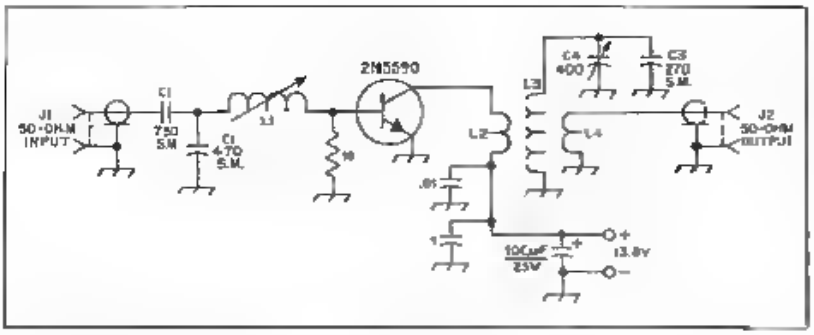


Fig. 1 — Schematic diagram and parts information for the K8EEG 40-meter amplifier. Capacitor values not otherwise marked are in pF. Some parts are numbered for text reference only. All grounds should be made directly to the transistor emitter strip. C1, C2 — Final values given; can be made variable as with C3-C4, for experimental purposes.

C4 — 400 pF mica dip trimmer. Small broadcast type capacitor suitable for low power application. See text.

L1 — 9 turns No. 22 enamel, close-wound on 1/4 inch dia. slug-tuned form.

L2 — 2.5 turns No. 22 enamel, close-wound on Amidon T-50-2 toroid core.

L3 — 13 turns, spaced to occupy entire core of L2.

L4 — 4.5 turns, spaced over 1/3 of core. In using the toroidal circuit for interstage coupling, make L1 1 to 2 turns for 10 to 40-ohm collector load impedance and 4 turns for 40 to 80 ohms.

If the amplifier is to be used with a separate exciter, as in this instance, the input network is designed and adjusted to match the low-base input impedance to 50 ohms, the usual output impedance of such an exciter. Where the amplifier is to be part of a transmitter, the collector circuit of the driver can be connected in place of J1. To provide for matching the capacitors C1 and C2 should be made variable in this case. A better way would be to make a toroidal matching transformer similar to L2-L3-L4, using slight alterations for this application given under Fig. 1. In the first case there are two unknowns present. The output capacitance of the driver and the input impedance of the amplifier base. This makes optimum adjustment rather complicated, since the output capacitance of the driver stage varies with its collector load impedance. With the tuned circuits in both stages, the driver can be optimized for 50 ohms and will work equally well when the amplifier is installed.

There are additional advantages. The tuned network will provide at least twice the harmonic rejection, and there will be much less loading of the previous stages by the final amplifier. The latter is very important in simple VFO-controlled transmitters, where pulling of the oscillator can result in considerable difference in frequency between the SPOT and OPERATE conditions.

The circuit used for the output network is a matter of personal preference. The double-link tank shown yielded an efficiency in excess of 50 percent at 7 MHz, so it was left in. In a 20-meter application the efficiency was about 40 percent. Conversion to the network described by Hayward (reference 2) brought the efficiency up to 62 percent.

Test Equipment

Three simple instruments, shown schematically in Fig. 2, were used in the development of the amplifier. A roughly calibrated wavemeter capable of tuning to the desired frequency and to its second harmonic, a power output indicator, and an impedance bridge. The wavemeter, Fig. 2A, was calibrated with the aid of a multiband transmitter.

The power-output meter, Fig. 2B, should be isolated from the transmitter and dummy load by shielding and RFC2. Actual output is obtained from the formula

$$P_o = \frac{V^2}{2R_1}$$

The meter is used to measure power output from a driver or amplifier stage during developmental work. Remember that it is not frequency sensitive. It will read combined fundamental and harmonic power, hence the need for the wavemeter.

The variable impedance bridge, Fig. 2C, is similar to one described by Hayward (reference 2) except that the diode is connected to the arm of a 1000-ohm variable control, instead of to the junction of two 470-ohm resistors. The control can be calibrated by connecting fixed resistors of known value across the output. Adjust the control for null, and mark down the resistance value used for that setting. When you want a circuit to look like, say, 70 ohms, you set the control to 70 and adjust the circuit for null. Parts placement is not critical, but it is wise to use short lengths of coaxial line in connecting the bridge into the circuit to be tested, and to ground both braids at the same point.

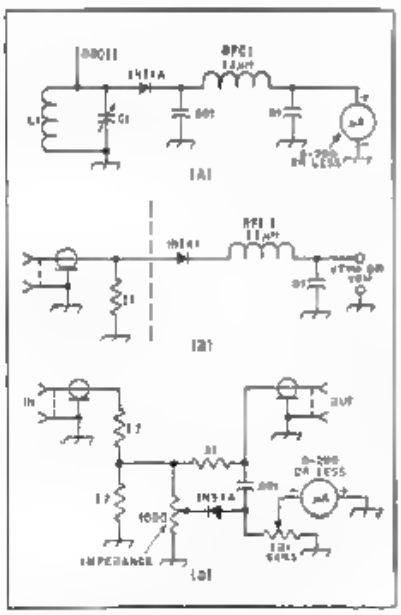


Fig. 2 — Simple test equipment used in optimizing the solid-state amplifier includes a wavemeter, A, a power-output indicator, B, and a variable impedance bridge, C. Values of L1 and C1 depend on the band being checked. Parts designations are for text reference.

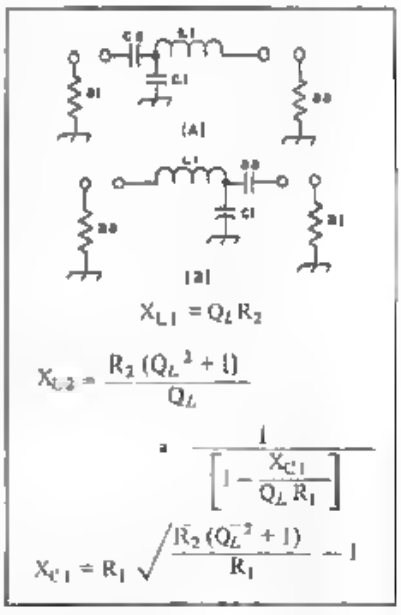


Fig. 3 — Basic circuit for use with Tables 1, 2 and 3. Circuit A, for Tables 1 and 2, shows the network for input matching. B is used in matching the amplifier to 50-ohm output. The formulas are for operating conditions other than those assumed in the tables.

If the bridge is to be used only between 50-ohm circuits, coaxial connectors will be suitable, as shown.

Construction and Testing

Armed with the above assumptions and test equipment, we can monitor several aspects of the circuit operation in the process of getting the amplifier to work properly. This is a rough duplication of the procedure followed in the manufacturer's laboratory in determining the performance characteristics of a device for given sets of conditions. These appear later on a data sheet. Our purpose is not quite the same, in that we are not looking for a set of "numbers." Rather, we seek to take into account automatically the actual characteristics of the device on *hand*, in achieving optimum operation for our application.

An experimental amplifier can be bread-boarded or built on a circuit board similar to the one shown. It is recommended that a single parallel-tuned circuit be used for the output side of the amplifier during developmental work on the input matching. It can be replaced when the work is completed. Calculated values for both input networks, and the output network, Fig. 3A and B, respectively, are given below for the hf bands.

Table 1
Input network, R1 = 50Ω, R2 = 5Ω, Q = 5.

	3.5	7	14	21	28 MHz
X_{L1} = 25Ω	1.25	0.63	0.29	0.2	0.18 μH
X_{C1} = 31Ω	1400	700	390	260	170 pF
X_{C2} = 641Ω	750	370	180	150	85 pF

Table 2
Input network connected to driver stage collector and load impedance of 70 ohms (1.25 watts at 12 volts dc), R1 = 5Ω, Q = 5.

		3.5	7	14	21	28 MHz
X_{L1}	40	1.25	0.63	0.29	0.2	0.18
X_{C2}	21	730	400	380	380	260
X_{C1}	64	[illegible]	370	180	150	85

Table 3
Output network, final collector impedance 8Ω, (10 watts output at 13.6 volts dc), 50Ω load. (From Motorola AN-267.)

		3.5	7	14	21	28 MHz
X_{L1}	40	2.0	0.95	0.49	0.3	0.23
X_{C1}	65	720	350	175	125	90
X_{C2}	89	530	260	140	90	65

The formulas given in Fig. 3 can be used to calculate approximate values, should the driver stage operate at a different power level or load impedance. C1, C2 and L1 should be variable, to allow for initial adjustments. Inexpensive broadcast-receiver capacitors, 365 pF, are ideal for tuning. Where higher capacitance is needed, fixed-value micas can be connected across the variables. A 40-meter amplifier is shown in Fig. 1 with component values arrived at by experiment, as described below.

Apply at least 500 mW of drive to the network through the impedance bridge. The network is adjusted for deepest null, first by C1, where the indication will be broad, then by C2, which gives a deeper null, and finally by L1. This is done with the wavemeter coupled to the final-amplifier tank, and the output meter connected to the tank as an indicating load. No dc voltage is applied to the amplifier thus far, as only the feed-through energy will be monitored at this point. With one watt of drive there should be 5 to 15 mW showing on the output meter, when the latter is tuned to the drive frequency. Remove the impedance bridge and repeak slightly for maximum feed-through indication.

Set the wavemeter to the second harmonic frequency. If the drive is clean and the circuits are properly tuned, there should be little or no output detectable at the harmonic frequency. Recheck tuning for minimum harmonic level, if any arrows. Optimum adjustment should give maximum fundamental output and rejection of harmonic output.

Apply collector voltage, with no drive. If the transistor is the balanced-emitter type, full collector voltage may be used. With other types it is well to start with about 70 percent of the maximum. De-couple the wavemeter, in anticipation of the 40-dB increase in power to be expected, and apply drive. Readjust both input and output networks for maximum output and minimum harmonic power. The wavemeter should be coupled to the lead going to the output meter for the latter check, as harmonic currents circulate in the output tank, and coupling to it will give an erroneous reading of harmonic level when the amplifier is running normally. Measure the dc input power and the rf output power and compute the efficiency which should be at least 40 percent. Substitute the double-tuned tank circuit for the simple parallel-tuned one, if the output is low.

If an external exciter is to drive the amplifier, no further adjustment is required, and the amplifier is ready for service. If you intend to connect the input network directly to the driver collector, the impedance bridge is set to the desired collector load impedance. Figure (70Ω for 1.25 W at 12 V), and adjustment is made for best match. Each of these steps monitors some aspect of circuit operation, using the actual components available, and gives assurance that optimum results are being obtained.

The amplifier shown in the photograph was adjusted by these methods and was ready for use, in the last hours before Field Day, in about a half hour after it was assembled. Running at reduced power, it gave a good account of itself on 40 meters the following day, using the exciter previously described by the writer.

References

[1] Johnson and Artis, "Fundamentals of Solid-State Power Amplifier Design," Part III, *QST*, April, 1971.

[2] Hanson, "How To Use RF Power Transistors," *Ham Radio*, January, 1970. Hayward, "Interest Power for the Solid-State Amplifier," *QST*, May, 1972. Lowe, "15 Watt Solid State Linear," *QST*, December 1971. Hejhall, "Broadband Solid-State Power Amplifiers," *QST*, March, 1972. Schultzing, "Transistorized Amplifier for 40 Meters," *73*, August, 1964. White, "Thermal Design of Transistor Circuits," *QST*, April 1972.

[3] Wass, "Multiband FET VFO QRP$_p$ Transmitter," *Ham Radio*, July, 1972.

From October 1978 QST, p 34:

On Solid-State PA Matching Networks

I would like to pass along some observations I've made which are of interest to builders of solid-state Class C hf transmitters. I have found that the use of such matching networks as the commonly recommended L and T, as well as any other network with an inductor or series LC as the input element,[1] will inevitably result in improper circuit operation. The circuit will exhibit poor collector efficiency, reduced output, or a high transistor failure rate, unless one of these conditions is met: (1) the output transistor is very rugged (in which case it won't fail, but the other conditions will remain); (2) the transistor output capacitance is 100 pF or higher; (3) a Zener diode is connected across the transistor (more about this later); or (4) the network is modified in a manner I will describe.

Let's see what causes the problem. Although there is an optimum resistive impedance for a transistor to "see" (approximately $V_{cc}^2/2P_o$),[2] the transistor does not present this resistive impedance. Rather, it acts much as a simple on-off switch. At the instant the transistor is turned off, current flowing through the L charges a capacitance into the circuit junction. The dominant circuit presented to this current is parallel resonant, with L being the network input inductor and C the transistor output capacitance. C_{ob}, C_{cb} is in parallel with many shunt capacitances. The shunt "rings" at its resonant frequency, which is not necessarily related to the operating frequency.

Fig 1 shows the schematic of a typical 40-meter, 2-watt-output amplifier. Fig 2A is a photo of the oscilloscope waveform at the collector of Q1. The presence of 70- or 50-MHz ringing at the collector may be readily seen. I was able to obtain this picture only because the particular transistor was exceptionally rugged — several devices were destroyed in the attempt. Although the condition could be detected with a voltmeter nudged loosely to the collector circuit, it can only be observed with the aid of a wide-bandwidth scope. The instrument used to obtain these photos has a 250-MHz bandwidth.

A photo of the waveform at the load is shown in Fig. 2B. Distortion may be reduced by filtering, but — assuming the transistor is not destroyed — collector efficiency will be less than optimum. Typical efficiency will be on the order of 30 to 60 percent, rather than the 70 to 80 percent obtained from a well-designed amplifier stage.

An advanced circuit-analysis computer program was used to investigate the circuit of Fig 1, assuming perfect inductor, capacitors, source, load and a good model of the 2N3866 transistor. The graphical results of this analysis are shown in Fig. 3. Because of the use of perfect components, frequency and amplitude of the simulated waveform vary slightly from the real waveform shown in Fig. 2. The striking similarity in Fig. 2 and the presence of ringing in the simulation verify that the phenomenon is not a spurious oscillation in the transistor, but is due to stray capacitance and inductance or poor circuit layout. *It is inherent in the use of this type of network!*

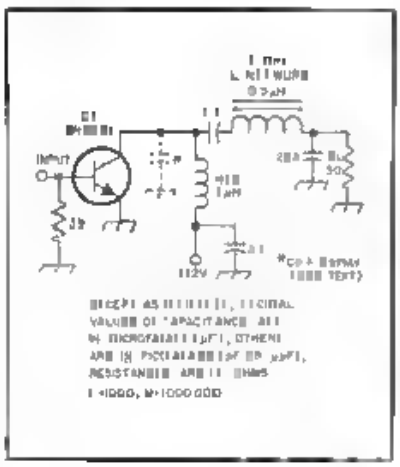


Fig 1 — Schematic diagram of the 7-MHz Class C amplifier used to examine the oscillation problem.

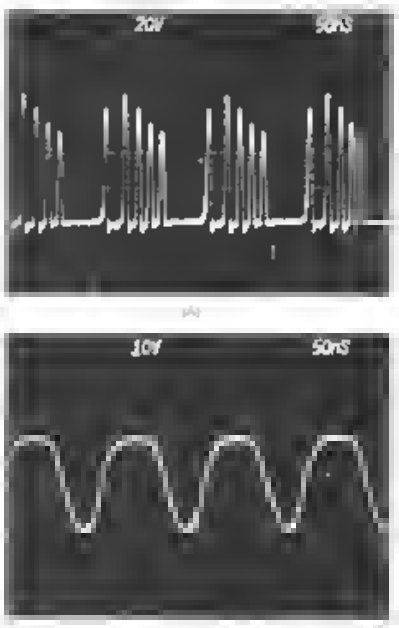


Fig 2 — Photos of the actual waveforms observed with the amplifier. Operation was observed with a high-speed oscilloscope. At A collector waveform; at B waveform at the output.

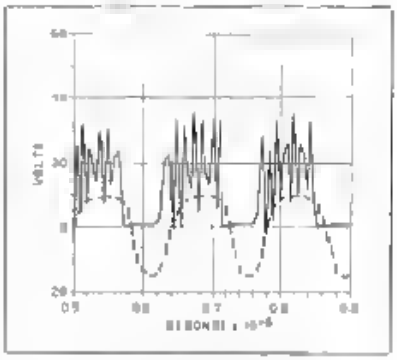


Fig 3 — When the circuit of Fig 1 was analyzed on a computer, the waveforms shown were predicted. The solid line indicates how the computer expected the waveform at the collector (tank circuit) output waveform is shown by the dotted line.

A separator connected from the collector to ground or, preferably, from collector to emitter, will solve the problem if it approximately resonates with the input inductor at the operating frequency. The capacitor will reduce collector-voltage swing to less than 30 volts with a 12-volt supply. The effect on the Q of common networks will be negligible and only slight readjustment of the variable capacitor(s) will restore the correct match.

A Zener diode connected across the collector will sometimes solve the problem, but not because of Zener action![3] A typical 33-volt, 1-watt Zener diode has a capacitance of 200 to 800 pF, depending on the amount of reverse bias. This is generally sufficient to prevent the ringing in the first place.

The latter has been necessarily brief but I hope it will enable the reader to take advantage of these matching networks without wondering — as I did for a long time — why sometimes they work and sometimes they don't. — *Ray W. Lewallen, W7EL, 5470 S.W. 152 Ave., Beaverton, OR 97005*

Footnotes

Strictly speaking, the Class C amplifiers used by amateurs may be better described as Class D as they are typically driven to saturation. In fact, this is the reason for the problem described here. However, such operation does allow high collector efficiency. For a more detailed discussion of this topic, see Sokal and Sokal, "Class E — A New Class of High-Efficiency Tuned Single-Ended Switching Power Amplifier," *IEEE Journal of Solid State Circuits*, Vol SC-10, No 3, June, 1975.

[1] Hayward and DeMaw, *Solid State Design for the Radio Amateur*, ARRL, 1977, pp 52-53.

[2] *The Radio Amateur's Handbook*, 54th Edition, 1977, ARRL, p 181.

[3] Hayward and DeMaw, p 24.

From January 1980 QST, p 55.

More on Solid-State PA Matching Networks

When I experimented with Class C bipolar transistor PAs about a decade ago I noticed the same phenomenon that W7EL reported in October 1978 *QST*. I also had the impression that the matching sections derived from shf ubf circuits are insufficient for rf power amplifiers.

The solution I adopted to cure this is somewhat different, however. I also place a capacitor (C_{pl}) from collector to ground, but, as a starting value, the reactance (X_{Cpl}) of this capacitor is about the same as the collector load resistance ($X_{Cp} = V_{cc}^2/2P_{dc}$, giving a loaded Q of 1.

In multiband transmitters, therefore, the collector rf choke remains the same and is chosen for the requirements of the lowest band in use (typically 80 µH for an 18-volt, 2-watt output PA). The capacitor C_p, is being switched from band to band together with the other tank-circuit components.

The loaded Q = 1 will stop the "ringing" phenomenon, but will also depress the efficiency to 50 to 60 percent. On the other hand, the PA becomes very insensitive to mismatch, even at full drive. For better efficiency and slightly reduced mismatch safety the loaded Q may be reduced to about 0.7. In higher power PAs where transistors with high internal capacitances are used, the Q may be reduced even further, depending on the efficiency obtainable.

I purposely place the capacitor C_{pl} from collector to ground, not to the emitter. As an additional measure for mismatch protection I recommend a small unbypassed emitter resistor, which causes a de drop of from 0.5 to 1 volt when the stage is fully switched. Placing C_{pl} to the emitter of this case would form a *regenerative* circuit.

When using rf silicon transistors (bipolar PAs) I further recommend adjusting V_{cc} to 0.5 $V_{CEO(sus)}$ or less for full-scale operation, because the breakdown voltages are lower at rf and lower figures are clear at shf.

Another trouble which I can throw up in tuned Class C transistor PAs is *frequency dividing*. The circuit may be tuned to, must have a configuration without any resonance at one half, one third etc. of the operating frequency. This is especially true if multisection pi networks are to be used for high harmonic suppression, or if a low-pass L section is to be added for antenna tuning.

Subharmonic resonances may be avoided by incorporating a parallel-resonant circuit (Fig 2) or a series-resonant circuit (Fig 3). The components marked * are selected by means of the band switch. I have been using the circuit of Fig 2 in my portable 5-band QRP transmitter (DB3553) PAs since 1969. It will match all random-length wires as well as coaxed antennas.

The circuit of Fig 3 will have somewhat better harmonic suppression, but the antenna must be safely grounded by means of a separate choke. Therefore, this scheme will be good for coaxed antennas and may take a rather high SWR, up to 3:1. — *Hans-Joachim Brandt, DJ2ZB, Lohenstenstrasse 7/b, 8000 Munich 60, Federal Republic of Germany*

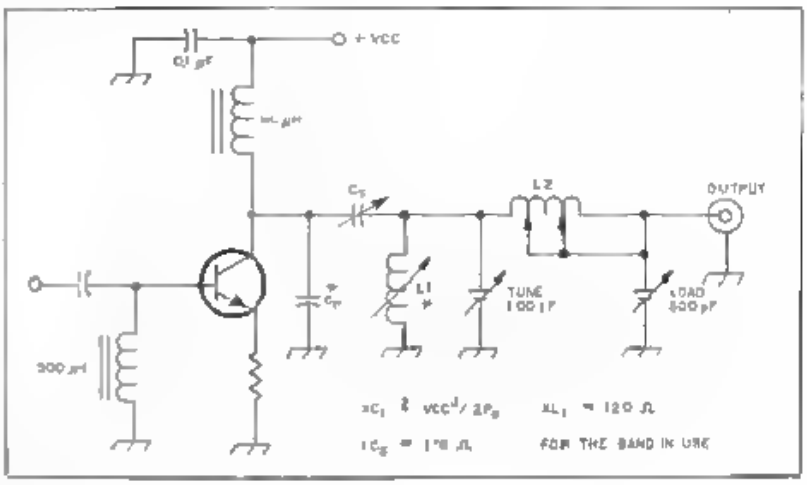


Fig 2 — Details of the parallel-resonant circuit

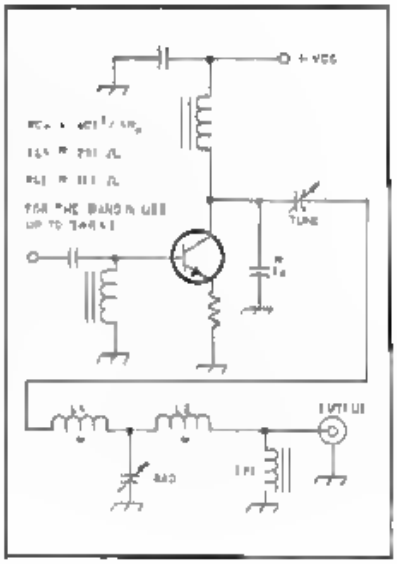
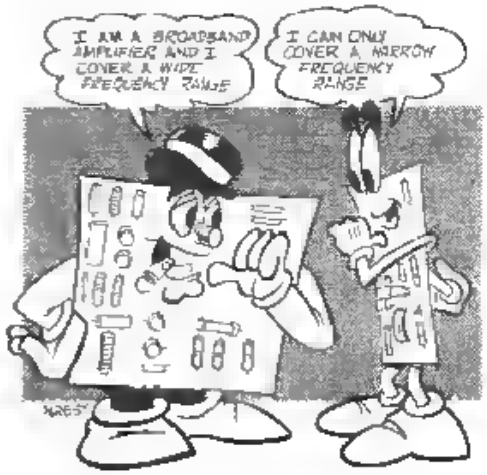


Fig 3 — Circuit for the series-resonant condition

From May 1984 QST, p 26:

Broadband and Narrow-Band Amplifiers



Narrow-band amplifiers have been around for many years, and most hams know how to design them. But, the broadband RF amplifier did not become popular until the semiconductor world bloomed. This article covers some practical aspects of both types.

By Doug DeMaw,* W1FB

Have you wondered what the difference may be between a narrow-band amplifier and a broadband one? Are all broadband amplifiers linear? Must they be linear? These are natural questions in the minds of most beginners in electronics, so we will try to provide simple answers.

If you work with transistors and RF circuits, it is likely that you will need to know something about how a broadband amplifier is designed, what to expect from it and how to build one for the job you have in mind. For the most part, these amplifiers are less prone to self-oscillation than are tuned, narrow-band styles of amplifier. The fundamental thought to keep in mind however, is that we must always trade some overall gain for increased bandwidth. If we can accept that trade-off, the major barrier will have been abolished.

Narrow-Band versus Broadband

The narrow-band amplifiers we use from day to day in our VFOs, receivers, converters and transmitters are *tuned* to some particular operating frequency. The tuned circuits are usually designed to yield a fairly high loaded Q (Q_L). The greater the circuit Q, the narrower the frequency response of the amplifier. Many applications require high Q and the attendant narrow bandwidth. Examples are VFOs, receiver front ends, transmitter tank circuits and filter circuits that contain an amplifier.

The narrow bandwidth is needed to reject unwanted signals above and below the desired operating frequency, and to prevent spurious energy from leaving the transmitter and reaching the antenna system. When broadband amplifiers are used in some of these more critical circuits, a filter of some kind must be used to obtain the desired spectral purity. By way of simple explanation, a broadband amplifier that has no filtering elements is merely an *untuned* amplifier. It will respond to a broad range of frequencies and, if designed well, should have relatively constant gain across that frequency range. An audio amplifier is but one example of a broadband amplifier.

Another advantage of the narrow-band circuit over the broadband type is that some circuits require minimum noise — as in the case of a receiver oscillator strip — and the high-Q tuned circuits greatly reduce the inherent noise output of the oscillator. High-performance receivers require "quiet" local oscillators in order to minimize "reciprocal mixing" in the mixer stage. Transmitter local oscillators should be similarly clean if we are to avoid broadcasting prohibitive amounts of broadband noise along with the desired signal output. Some commercial early-day solid state transmitters were very offensive in terms of transmitted wideband noise.

Fig. 1 shows examples of narrow band and broadband amplifiers in some simplified circuits. Illustration A shows a conventional small-signal RF amplifier with tuned circuits at the input and output. This is typical of what we may find at the input of a receiver. The high-Q tuned circuits or resonators restrict the frequency response for a given setting of C1 and C2. For this reason we will call our circuit a narrow-band amplifier.

Although the circuit at B of Fig. 1 is an oscillator, it is in reality a form of amplifier. For an oscillator to work as such it must be designed as an amplifier. Some of the output energy is fed back to the input terminal to cause oscillation. Again we have a high-Q tuned circuit (C3, C4 and L1), which restricts the bandwidth of the circuit in accordance with the particular setting of C3. Owing to our use of some of the output power as feedback, this type of amplifier is not as efficient as is the circuit in Fig. 1A.

Fig. 1C contains an example of a broadband amplifier for RF use. It operates linearly because it is biased for class A. T1 is a broadband transformer that can be used to match the amplifier impedance to that of the load by virtue of the transformer turns ratio. Note that T1 is untuned; hence the bandwidth.

A class-A linear broadband amplifier

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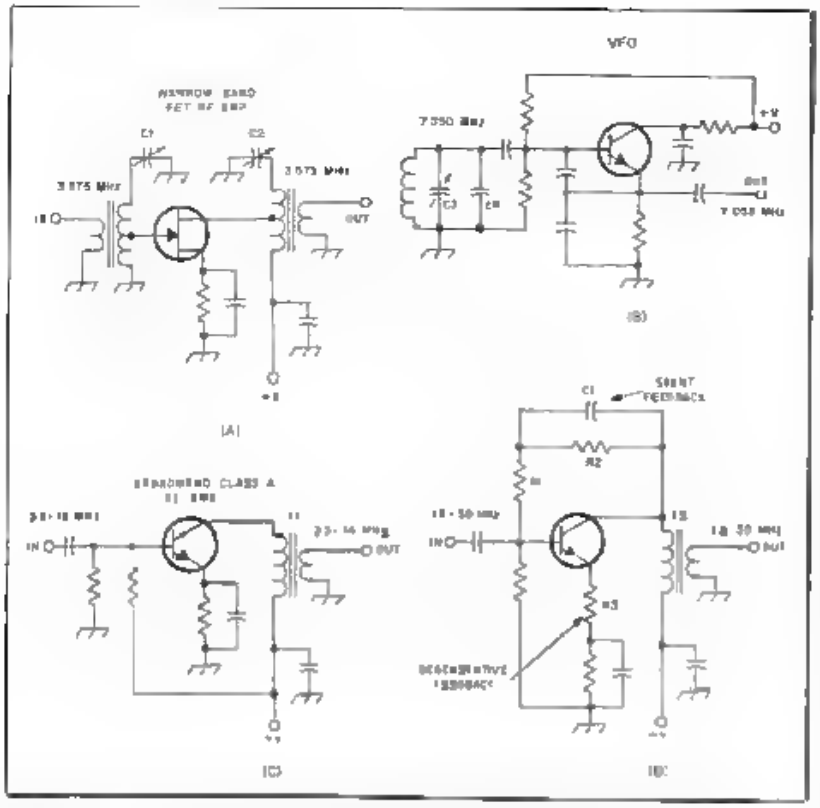

Fig. 1 — The diagrams at A and B illustrate narrow-band amplifiers. The VFO circuit is still a form of amplifier since its output power (in part) is fed back to the input for the purpose of causing oscillation. A simple Class-A broadband amplifier without feedback is seen at C. A fed back Class A broadband amplifier is shown at D. It uses a combination of shunt and degenerative feedback (see text).

with feedback is shown at D of Fig. 1. Here, we have intentionally introduced feedback by means of resistive divider R1/R2 and C1. Degenerative feedback is provided by means of the unbypassed emitter resistor, R3. This type of amplifier has considerable bandwidth. The shunt feedback stabilizes the current gain of the stage while decreasing the input and output resistance of the amplifier. The emitter degeneration helps stabilize the transistor voltage gain, and it increases the input impedance of the transistor. The increase is approximately proportional to the transistor beta. A specific treatment of feedback applied to broadband amplifiers is contained in *Solid State Design for the Radio Amateur*, available from ARRL.

Amplifiers with feedback are used not only for low-power circuits, but are practically the order of the day for high-power solid-state RF amplifiers. A circuit for a broadband, fed-back linear amplifier is provided in Fig. 2. Since this diagram is purely for illustrative purposes, no component values are assigned.

Assume that the circuit is capable of delivering 100 W of output from 1.8 to 29.9 MHz. Shunt feedback is made possible by the networks that contain R1, R2, R3, R4, C1 and C2. Here, we are applying negative feedback between the collectors and bases. Were we to use positive feedback, as in the case of oscillators, the amplifier would "take off" in a spasm of self-oscillation. *Positive* feedback is of the same phase as the input energy, whereas *negative* feedback is approximately 180 degrees out of phase with the input signal. This relationship is important to remember. An absolute 180-degree phase shift is difficult to realize when working with transistors, owing to some inherent phase shift as the signal current passes through the semiconductor material.

T1 and T2 of Fig. 2 are broadband transformers whose frequency response, if they are designed well, is reasonably flat across the 1.8-30 MHz range. Generally, ferrite core material of 800 to 950 effective permeability (μ_e) is used for high-frequency broadband amplifiers. This is a no. 43 material when ordering from Amidon Associates or Fair-Rite Corp.[1] Palomar Engineers and RadioKit also supply cores of the no. 43 variety. Core permeabilities of 125 and 40 are commonly used for VHF broadband transformers.

Broadband transformers work like this: As the operating frequency is increased, the core material becomes less and less effec-

[1]Notes appear at end of article.

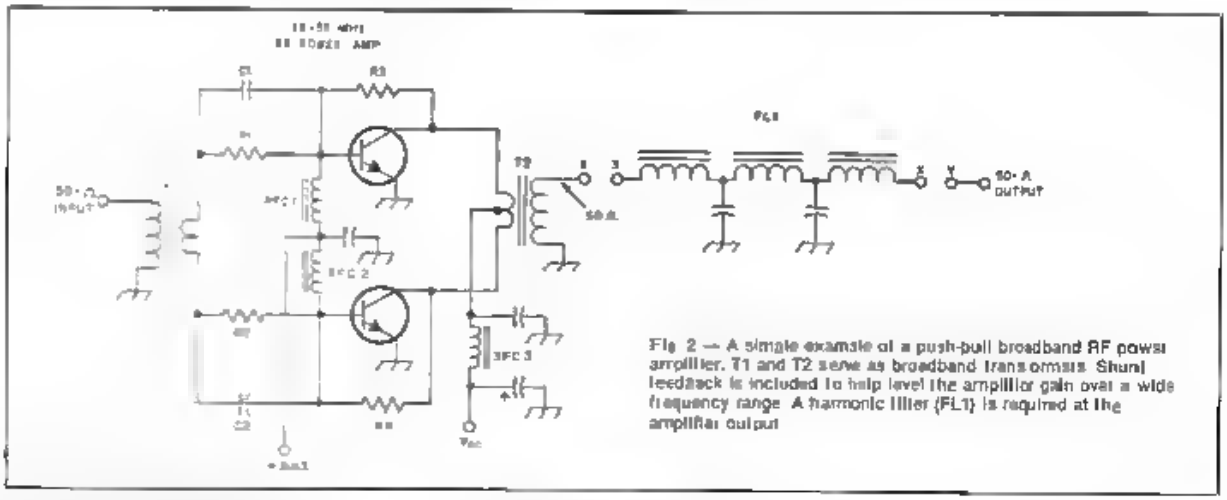

Fig. 2 — A simple example of a push-pull broadband RF power amplifier. T1 and T2 serve as broadband transformers. Shunt feedback is included to help level the amplifier gain over a wide frequency range. A harmonic filter (FL1) is required at the amplifier output.

tive in the circuit. At the low-frequency end of our transformer range, the core does its job and increases the inductance of the windings (necessary). At the high end of the transformer performance range, the core becomes essentially "not there" as far as the windings are concerned. This enables us to obtain a substantial bandwidth that would be impossible with coreless transformers. A suitable rule of thumb for transformer design is to make the inductive reactance of the smallest winding approximately four times the load impedance. Hence, if the base of a transistor amplifier exhibited a 10 ohm impedance, the broadband-transformer winding that we connect to the base should have sufficient inductance to have a reactance of 40 ohms or slightly greater. If not, the low impedance of the winding would shunt part of the driving power to ground and could cause an SWR condition.

Let's assume that our amplifier is operating at 7.1 MHz. The base impedance of the transistor with drive applied is 12 ohms. How much winding inductance would we need for the transformer secondary? The standard equation for inductance would be used:

$$L(\mu H) = \frac{X_L}{2\pi f(MHz)} \qquad \text{(Eq. 1)}$$

So, with an X_L of 4 times 12, we would obtain the following answer:

$$L(\mu H) = \frac{48}{6.28 \times 7.1} = 1.07 \qquad \text{(Eq. 2)}$$

The required number of turns can be calculated from

$$\text{Turns} = 100\sqrt{L(\mu H)/A_L} \qquad \text{(Eq. 3)}$$

where A_L is the number provided for each type of core by the vendor or manufacturer. Each core relative to its cross-sectional area and the core material, has a specific A_L factor. The Amidon Associates catalog contains such data, as does a book concerning magnetic cores.[3]

I don't want to mislead you into thinking that broadband amplifier design is a snap. There are many subtleties involved, and considerable study of the pertinent literature is important before launching one's own project from scratch. Motorola Semiconductor Company has a wealth of useful data in its book on power semiconductors, inclusive of application notes on transformer and broadband amplifier design.[4]

But, let's return to Fig. 2 and learn a bit more about what is going on. T2, the output transformer, serves also as an impedance-matching device. The inductances in the transformer windings are based also on a 4:1 rule, respective to the collector impedance. This impedance can be calculated closely from $Z = V_{cc}^2/2P_o$ ohms. Eq. 1 is then applied. FL1 is a harmonic filter, and is a low-pass type. A switch can be inserted at points X and Y to permit band switching of the low-pass filters. This is standard procedure in commercial equipment. For single-band use, a jumper can be placed across X and Y.

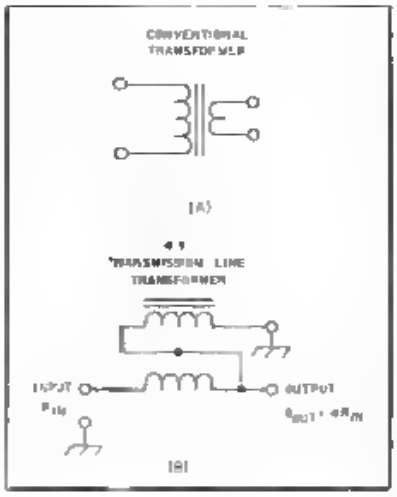

Fig. 3 — Examples of conventional and transmission-line transformers. See text for additional information.

It is important in all broadband amplifiers to minimize the stray capacitive and inductive reactances. These parasitic quantities of L and C have a marked effect on the amplifier performance as the operating frequency is increased. In other words, unwanted capacitive and inductive reactance will limit the upper frequency response of the circuit. An improperly designed broadband transformer will degrade the performance in a like manner.

If we aim to minimize the presence of stray reactance, we must use large or very short circuit-board strips. This will reduce the effective inductance of the PC-board foils. These copper strips should also be as direct as possible. Similarly, the connecting leads of resistors and capacitors must be held to a minimum length. Many amplifiers contain chip resistors and capacitors to keep stray inductance and capacitance to a minimum. These components are supplied without leads or "pigtails." They are soldered directly to the PC board foils. They are practically a requisite at the upper end of the HF range and higher, but they are more costly than are silver-mica or disc-ceramic capacitors.

Conventional or Transmission-Line Transformers?

I'm sure you've heard designers speak of "conventional" and "transmission-line" transformers. The so-called conventional transformer is built along the lines of an audio or power transformer. That is, it has a core and separate windings, as in Fig. 3A. The transmission-line transformer, on the other hand, has bifilar, trifilar or quadrifilar windings that are placed on the core in parallel, or they may be twisted together beforehand. In this case, each winding conductor is the same length. The windings function as short lengths of transmission line, and the impedance is generally 25 ohms. Either style of transformer can be used in a broadband amplifier, or as a matching transformer in other types of circuits, such as antennas.

The conventional transformer is considered less efficient than the other type, but it enables us to obtain nearly any turns ratio we desire. The transmission-line transformer (Fig. 3B) yields only specific integers of transformation, such as 4:1, 9:1, etc. Furthermore, we can find ourselves rather frustrated when trying to hook up a multiwire transmission line transformer, especially if the same size and color of wire is used for the windings. Many engineers use enameled wire of various colors to avoid this problem. Green, red and brown wire is often used. You can solve the problem by dipping the wires in different colors of paint before using them. I have had good results by spraying the wires with fast-drying paint.

A Handy Broadband Amplifier

Many times we find ourselves in need of a little extra "push" when working with a scope or frequency counter. Perhaps the sampling point in the circuit has insufficient signal voltage to trigger our frequency counter or cause ample deflection on the face of the scope tube. A broadband amplifier is useful at such times to give that weak signal the needed boost.

Our workshop project this month is shown schematically in Fig. 4. It is patterned along the lines of a broadband amplifier designed by Hayward, W7ZOI. His design did not use transformers and there was no high-level stage at the far end of the amplifier strip, but the feedback networks are similar to his. The particulars of the general design are given in the text of *Solid State Design for the Radio Amateur*, referenced earlier in this article.

CATV transistors are used to ensure good bandwidth (1.2 GHz f_T) and linearity. Each stage is biased for linear Class-A operation. A combination of shunt and degenerative feedback is used throughout the circuit. The input of each amplifier is roughly 50 ohms, and each output is approximately 200 ohms with the values given. Amplifier stability is excellent, even when there is no termination at the input and output ports. Circuit boards and parts kits for this circuit are available.[4]

The bandwidth is flat from 400 kHz to 34 MHz (within 1 dB). I measured the overall gain as 41 dB. The maximum-acceptable output, in terms of distortion, is 0.25 W. The circuit draws 90 mA of current with a supply voltage of +13.

Owing to the linearity and bandwidth of the circuit in Fig. 4, it is ideal as a drop-in

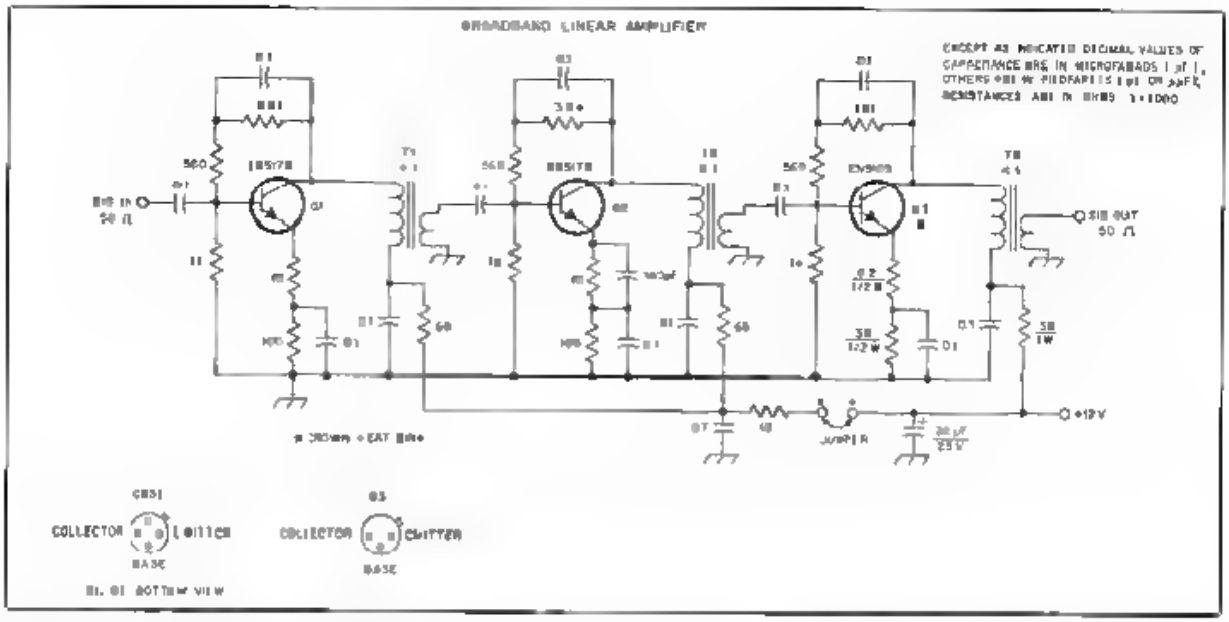

Fig. 4 — A practical circuit for a broadband linear-amplifier strip. This can be used as an instrument amplifier, a low-level RF strip in a transmitter or as part of a receiving loop preamplifier. Resistors are ¼-W carbon-composition unless otherwise noted. The polarized capacitor is tantalum or electrolytic. All others are chip-style or disc-ceramic with short leads. T1 and T2 contain 15 primary turns of no. 28 enameled wire on an Amidon FT37-43 toroid core. The secondary windings consist of seven turns of no. 28 enameled wire. T3 uses an Amidon FT50-43 toroid core with 12 primary turns of no. 26 enameled wire. The secondary of T3 contains six turns of no. 26 wire.

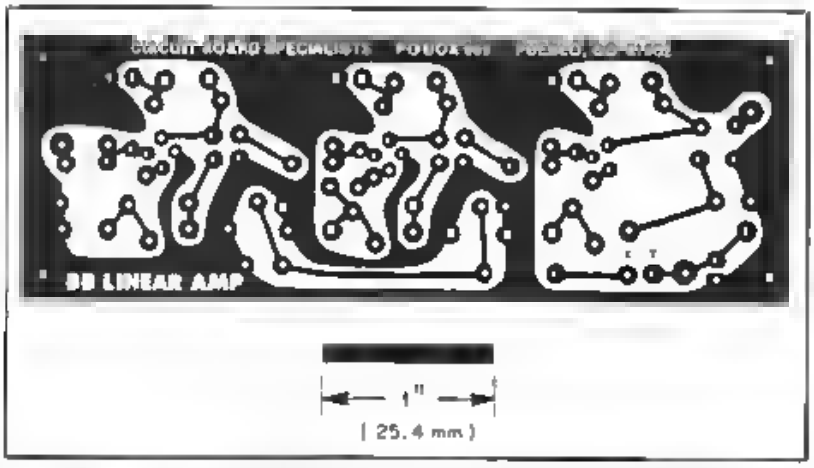

Fig. 5 — Circuit-board etching pattern for the broadband amplifier of Fig. 4. The pattern is shown full size from the foil side of the board. Black areas represent unetched copper foil.

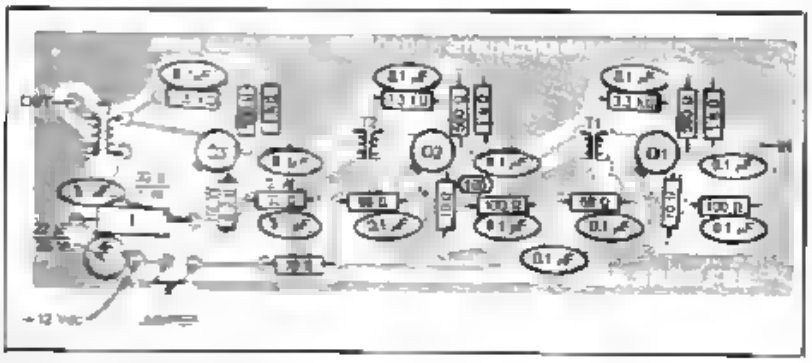

Fig. 6 — Parts-placement guide for the broadband amplifier of Fig. 4.

unit for an HF-band CW or SSB transmitter. It can be used at the low-level section of such a transmitter. I wish to caution you, however, that it should not be used for QRP operation unless a suitable harmonic filter is placed between the amplifier output and the antenna. A half-wave style of filter should be suitable if you want to try your hand at low-power operation.

Terminals X and Y on the circuit board are available for use as a standby point, or for CW keying. If a keying line is attached at X and Y, be sure to include a shaping network so that your signal won't sound clicky.

This amplifier can be used also as a preamplifier for loop antennas. A step attenuator can be inserted at the output of the amplifier to control the gain. If you choose to use this circuit in such a manner, a low-noise preamplifier should precede Q1 of Fig. 4. I find that a JFET stage is suitable for this purpose. Owing to the small signal that a receiving loop provides, the preamplifier (even at 1.8 MHz) must be a low-noise type. If not, you will enjoy listening to "pop-corn" noise along with the DX signals! Q1 does not have a low enough noise figure for satisfactory weak-signal reception.

Construction

If you choose to make your own PC board for this project, try to keep all stages in a straight line. Keep the PC-board foils short and direct. Minimize the lead length of each capacitor and resistor. Make sure the transistors are seated close to the PC board in order to keep their leads as short

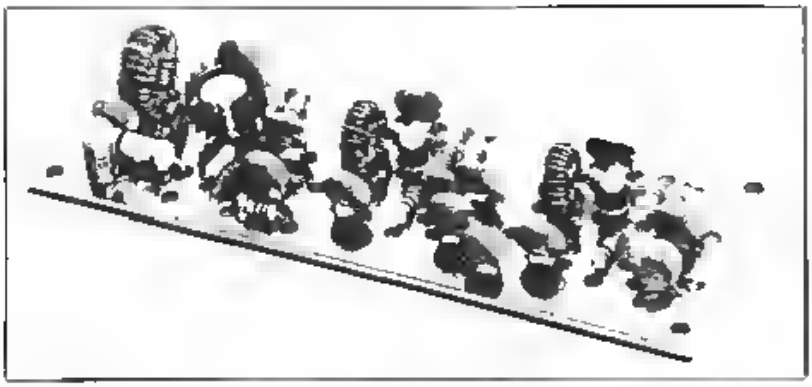

Fig. 7 — The assembled broadband amplifier. Note that in-line layout is used.

as possible. A crown heat sink is needed on Q3, the 2N5109. A coating of silicone grease should be applied to the transistor cap before installing the heat sink. Double-sided PC board is recommended in the interest of stability. Fig. 6 shows the parts placement for the circuit board, as seen from the component side. A scale template of the PC board pattern is provided in Fig. 5. Fig. 7 is a photograph of the assembled amplifier.

Some Final Remarks

I hope you have learned the basics about narrow-band and broadband amplifiers. Certainly, we've only scratched the outer layer of the subject. A thorough treatment would require several *QST* installments. Our purpose this time is to explain the difference between amplifier types, and to provide a project that would enable you to try your hand at broadband amplifier construction and use.

A broadband amplifier can be built for Class A, B or C service just as narrow-band amplifiers can. The advantage of broadband designs is, in retrospect, to obtain a wide frequency response with relatively flat gain. This helps us to design circuits that do not require band-switching provisions. In other words, it simplifies the design of a multiband transmitter. But, as an instrumentation amplifier, the circuit of Fig. 4 has a great many advantages around the workshop. Good luck with your project.

Notes

[1]Amidon Associates, 12011 Otsego St., N. Hollywood, CA 91607 [catalog available. Fair-Rite Products Corp., 1 Commercial Row, Wallkill, NY 12589. See *QST* ads for Palomar and RadioKit.

[2]D. DeMaw, *Ferromagnetic Core Design & Applications Handbook*, pp. 0-13-114068-1 (Englewood Cliffs, NJ: Prentice-Hall, Inc.)

[3]*Motorola RF Data Manual*, Motorola Semiconductor Products, Inc., P.O. Box 20912, Phoenix, AZ 85036.

[4]Circuit Board Specialists, P.O. Box 969, Pueblo CO 81002, tel. 303-542-5081.

For updated supplier addresses see ARRL Parts Suppliers List in Chapter 2.

From September 1984 QST, p 19

Electronic Switching and How It Works

Replace those old-fashioned toggle switches with up-to-date diodes and transistors and you'll have simpler, less expensive and less cumbersome circuits.

By Doug DeMaw,* W1FB

What could be more ordinary than a switch? True, they are not very spectacular devices, but few circuits can be made to function without some type of switch — mechanical or electronic. Electronic switching is not new, but the state of the switching art has moved forward in grand style since semiconductors became as common as patent medicines. Furthermore, the cost of a solid-state switching device (diode, IC or transistor) is generally less than that of a comparable mechanical unit, such as a toggle switch.

Substantial levels of ac and dc power can now be switched by means of large diodes, power FETs, Triacs, and the like. Also, relatively high potentials can be accommodated safely by some rather small semiconductor components. At the dawning of our solid-state era, we were able to switch low amounts of signal and dc, and at fairly low voltage levels. It seemed in those days that the technology was not going to offer much promise toward replacing cumbersome manual switches with tiny diodes or transistors, but the trend today is clearly toward semiconductor switches. Amateurs can take advantage of the many options presented by solid-state switches, so let's examine a few basic concepts and see how we can develop practical circuits that use diode and transistor switches. First, in the interests of accuracy, let's look at the shortcomings of electronic switches.

Some Limitations

There is no magic in the electronic-switching art. In other words, we can't achieve everything that mechanical switching offers. But, we can come close to realizing the concept of universal replacement of mechanical switching components. What are the trade-offs? First, high-power RF switching is still a tough assignment with present-day low-cost transistors or diodes. Second on the list of not-so-nice features is that large solid-state switches need heat sinks of substantial size, and they may also call for cooling fans. This results in mass and expense that is not acceptable for amateur projects. High-power switches can become larger and more costly when using semiconductors. Number 3 on my list is the inherent internal resistance of most solid-state switches. It is seldom possible to have a zero resistance through a semiconductor switching device. Although the resistance of such a semiconductor junction in the ON mode may be only a fraction of an ohm, it can be enough to cause a problem. Some semiconductor switches have internal resistances greater than an ohm when activated. This becomes a source of difficulty as high power (heating and voltage drop by virtue of the I^2R rule), and in certain types of switching circuits it means that *complete* switching is not possible. For example, the internal resistance of a power FET is specified as RDS (resistance from drain to source) when it is switched to the ON state. This will vary with the device, and can range from 0.5 ohm to a few ohms, depending on the particular FET chosen. Well-designed mechanical switches, on the other hand, will exhibit a nearly zero-resistance condition between the contacts.

How else might we vilify the solid-state switch? Well, we should mention that input-output isolation is seldom of the magnitude that we can obtain with a suitable mechanical switch. This is caused by the semiconductor internal resistance and capacitance. It is a concern mainly when we wish to use a semiconductor to switch a signal line; for dc applications it is not a matter of importance. Finally, in many circuits that contain electronic switches, we need to actuate them by means of a mechanical switch. However, it is often practical to control dozens of electronic switches simultaneously with a single SPST mechanical switch, and therein lies the advantage!

Some Basics

Fig. 1 illustrates the fundamental principle of mechanical and solid-state switching. Assume we wanted to apply dc to a specific module. Example A shows the mechanical means to do this. Circuit B relies on a bipolar transistor to switch the dc on and off. The dashed lines show that the base of Q1 must be grounded to activate the circuit. This can be managed by the use of a mechanical switch, or by triggering Q1 with another semiconductor switch elsewhere in the system. The use of a PNP transistor permits application of the +12 V to the emitter, and also enables us to turn Q1 on by grounding the base through R1. If we used an NPN transistor at Q1, we would need to apply +12 V at R1 in order to saturate (switch) the transistor. Also, the +12 V of operating potential would have to be fed to the collector rather than to the emitter, as shown.

By grounding the PNP transistor base, or through applying +12 V to R1 of an NPN device, we are providing what is called *forward bias*. This causes the transistor to conduct heavily, which makes it perform the switching function. Too much current, caused by excessive base-emitter voltage, can destroy the transistor. Therefore, a series resistor is used (R1). Too little forward bias, conversely, will prevent the transistor from saturating com-

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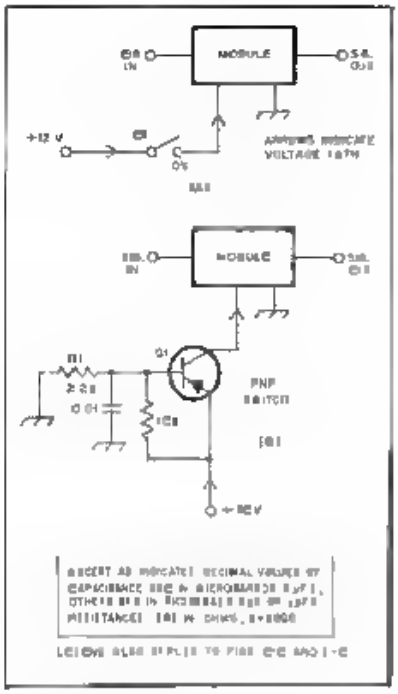


Fig. 1 — Comparison between a simple mechanical switch and an equivalent solid-state type.

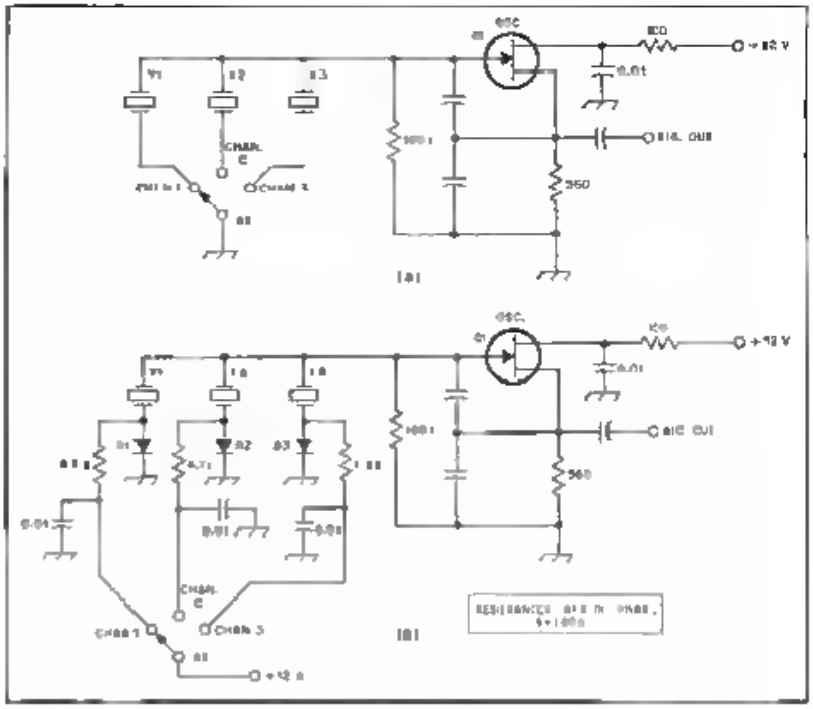


Fig. 2 — Diodes may be used in place of a mechanical switch to select oscillator crystals, as shown at B.

pletely. This will result in partial operating voltage reaching the module of Fig. 1B.

A comparison between mechanical and diode switching is offered for your study in Fig. 2. Circuit example A illustrates the old way of selecting crystals in a multi-frequency oscillator. This method is acceptable if the switch leads are very short or if the crystals are mounted directly on the switch. It is necessary to always keep RF leads as short as possible to prevent impairment of the performance.

Fig. 2B demonstrates the use of diodes in place of S1 of Fig. 2A. This enables us to locate the selector switch a great distance away from the crystals, and the connecting leads will carry dc rather than RF. The diodes offer a practical convenience, and the same control switch may be used to actuate many solid-state switches elsewhere in the overall equipment when the crystals are selected one by one. The 4.7-kΩ resistors near D1, D2 and D3 limit the current that flows through the diode junctions. They also serve as RF chokes in the switching lines to S1. In this low-power circuit, we can safely use inexpensive 1N914 high-speed switching diodes. They are suitable into the microwave region. When dc is routed to a diode through S1, it becomes forward biased and conducts, thereby completing the circuit between the bottom end of the crystal and ground. Diodes can be used in a similar manner to complete various RF circuit paths. They are often used in *series* with signal lines.

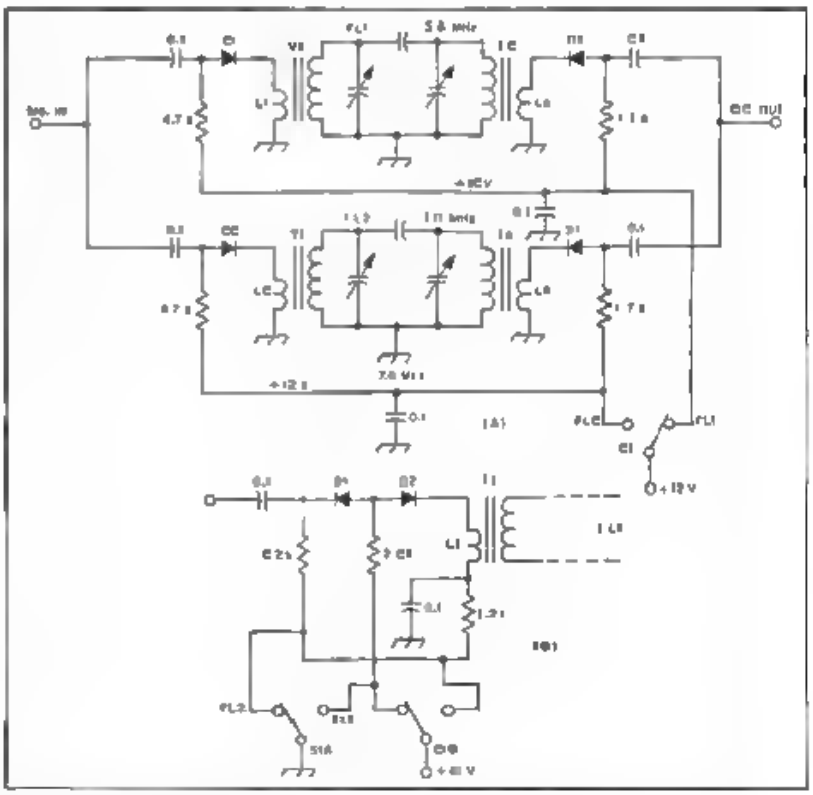


Fig. 3 — Series-diode switching is illustrated here. Example A shows how a pair of band pass filters could be selected by means of diode switching. A better technique is shown at B (see text).

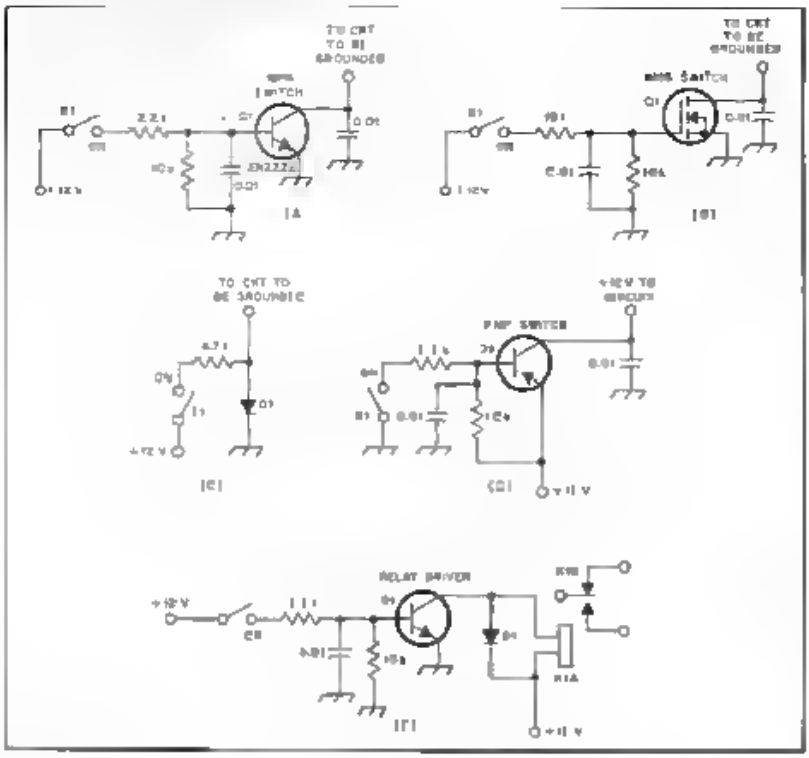


Fig 4 — Various semiconductor switches (see text).

Examples of series diode switching are given in Fig. 3. The circuit at A shows a typical arrangement in which we might use diodes to select band-pass filters. The input and output ends of each filter are connected to switching diodes to permit electronic insertion or removal of the desired filter. This circuit will function as shown, but it is a simplification of diode switching, for the purpose of making the example less difficult to understand. Each diode has a 4.7-kΩ resistor in series with the related +12-V line to limit the junction current and to function as an RF choke. If the resistors were not used as chokes, the input and output signal to and from the filters would be lost to ground through the +12-V line. Each diode obtains its dc ground return through the tuned circuit windings (L1, L2, L3 and L4). Small diodes of the 1N914 variety work well in this circuit.

A better way to employ diodes for series switching is shown in Fig. 3B. Here we have two diodes in a back-to-back arrangement. This circuit would be used at each end of FL1 and FL2 of Fig. 3A. The advantage of using two diodes is better isolation of the filters. Forward bias is applied to the diode anodes via S1 when a filter is selected. The unused filter (not shown) is well isolated from the signal line because reverse bias (+12 V applied to the diode cathodes) is switched to the dormant diodes to prevent any conduction caused by RF energy that may be present. When S1 of Fig. 3B is set for FL1 use, D1 and D2 are turned on and the signal path is completed. When S1 is changed for use of FL2, D1 and D2 are reverse biased to turn them off. Additional ground and +12-V lines are also connected to the contacts of FL1 to control the remaining six diodes that would be used for two filters of the type indicated in Fig. 3A. One DPDT switch would be used to control all eight diodes.

Here we see the advantage of solid-state switching, for if a mechanical switch were used, it would require four poles with two positions each. Also, the lead lengths from the filters to the switch section(s) could be prohibitive, and the input/output isolation of the filters could be poor because of signal leakage across the switch section(s). RF chokes can be used in place of the resistors if desired, but there should still be a current-limiting resistor in the dc line to the diodes.

Basic Shunt Switches

Fig. 4 contains a number of examples that show how we may use various semiconductor devices as shunt switches. S1 in each case represents a mechanical switch or CW key that turns on the electronic switch. As we learned earlier, function can be effected also by other electronic switches in the overall circuit. For example, AGC voltage or rectified speech energy could be used to actuate a semiconductor switch, depending on the application for the switch; the possibilities are virtually without limit.

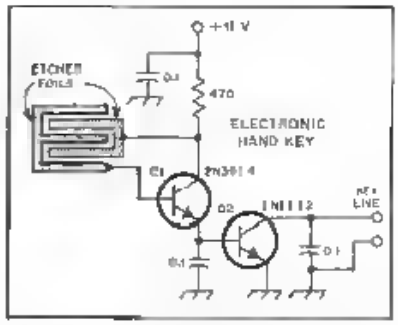


Fig 5 — Practical circuit for an electronic send key. The etched foil grid at the left is the keying contact. Transmitter or code-practice oscillator keying occurs when the operator places his or her finger on the copper grid. The resistor is in ohms, and the capacitors are in microfarads.

Circuit A of Fig. 4 is an NPN transistor switch. A positive voltage is required at the base of Q1 in order to turn it on. The 10-kΩ resistor from the base to ground is used to minimize transistor leakage current when the switch is in the OFF mode. This circuit might be used to key a driver stage in a solid-state CW transmitter. If so, the collector of Q1 would be hooked to the emitter of the keyed stage. S1 would be the CW key, or the keying line from a keyer.

Fig. 4B shows a similar circuit, but with a power FET switch. Since we have an enhancement-mode FET in our circuit, the transistor requires a forward gate bias to turn the device on and make it switch.

A simple diode switch is shown at C of Fig. 4. To the right is a PNP transistor switch (D). The base must be grounded through the 2.2-kΩ resistor to ensure turn-on. A 10-kΩ resistor is connected from base to the +12-V line to help cut off the transistor in the OFF mode. This resistor can be eliminated in the circuits of Fig. 4A, 4D and 4E if two 1N914 diodes are connected in series with the emitter leads. The diode junctions will reverse bias the transistors by approximately 1.4 V. The resistor or biasing diodes are especially important in the circuit of Fig. 4E — the relay driver. Without reverse bias, the relay may remain energized after the transistor is turned off. This is because a small amount of leakage or idling current will remain, and it may

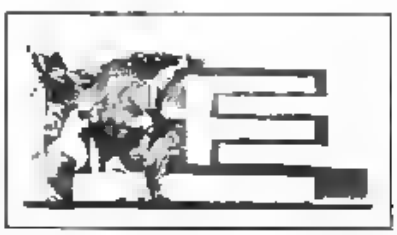


Fig. 6 — The semi-model key shown in Fig. 5.

be ample to keep the relay closed once it has been energized. D1 of Fig. 4E is used to clamp voltage spikes that occur when the field coil of K1 collapses at turn-off. Such spikes, if allowed to exist, could follow the +12-V line and damage semiconductor devices elsewhere in the circuit.

An Electronic Straight Key

So that we may understand how two transistors can be arranged to work in concert as dc switches, let's look at Fig. 5. Q1 is made to turn on when we place our finger on the circuit-board foil at the left. The resistance of our skin completes the bias circuit for the base of Q1. This actuates Q1, which in turn fires Q2, the keying transistor. When Q1 switches on, dc voltage appears at the emitter. At this moment, the positive potential also reaches the base of Q2, causing it to switch to the ON state. This closes the keying line to our transmitter.

There is no current-limiting resistor at the base of Q2 because there will not be a prohibitive voltage level coming from Q1. This is because the contact resistance of our fingers is sufficiently high to prevent Q1 from completely saturating. Hence, the output voltage from the Q1 emitter will be low enough for safe operation of Q2. A husky transistor can be used at Q2 if the key-line current warrants a transistor with a dissipation rating greater than that of the 2N2222. A 2N2102 (or equiv.) would be a good choice.

A photograph of a crude test model of the hand key is presented in Fig. 6. The copper grid is etched as shown, and isolating pads are used to contain the transistors and related parts. A three-circuit keying is needed to accommodate the +12-V, keying and ground leads. This key will operate satisfactorily from a 9-V transistor-radio battery as well. Three bypass capacitors are used to help keep unwanted RF energy from affecting transistor performance.

Fig. 7 shows the driver and PA stages of a simple QRP CW transmitter. The key from Fig. 5 could be used to operate the dc switch, Q3, of Fig. 7. When our finger is placed on the copper grid of the key Q3 will turn on. This action will permit the flow of dc to driver Q1, thereby keying our transmitter.

D1 and D2 of Fig. 7, 1N914 small-signal diodes, are used as a TR (transmit-receive) switch. This circuit was introduced by Wes Hayward, W7ZOI, and has been used for QSK (full break-in) in many of his QRP rigs. In this example, the diodes are turned on by RF energy from the collector of Q2 when Q3 is actuated. Some of the RF voltage is sampled by C1 and is routed to the diodes. During transmit, D1 and D2 are shorted to ground, thereby protecting the receiver input circuit. The diode conduction threshold is roughly 0.7 V. As a result, there will be a 0.7 V RF potential appearing at the receiver input circuit. This is not a great enough voltage to cause harm to the receiver. I have applied that amount many times to the input line of my FT101E, FT301D and FT102 transceivers, and no damage resulted. Greater details of this type of TR circuit are given in the League's book, *Solid State Design for the Radio Amateur*.

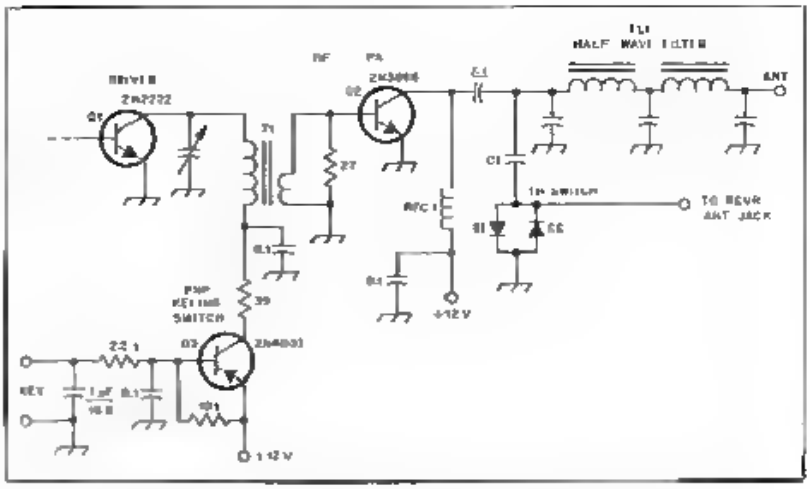


Fig. 7 — Example of a transistor keying switch (Q3) and two diodes (D1 and D2) used as TR switches for QSK operation.

One of the penalties for using the simple TR circuit of Fig. 7 is a loss in received signal (about 6 dB, from my experience). This is because C1 must be relatively small in value to prevent it from affecting the design of the output network of Q2. I use a capacitive reactance of 400 for C1. Thus, at 7 MHz, we would have a 56-pF capacitor at C1. The signal loss can be corrected by inserting a low-gain RF amplifier between the TR switch and the receiver input line. Fig. 8 shows how this might be done. A grounded-gate JFET, such as an MPF102, would serve nicely at Q5.

A More Elaborate Switching Circuit

An illustrative transmitter circuit is presented in Fig. 9. The arrangement for Q4, Q5 and Q6 is one I developed for personal use with a few QRP rigs up to 3 W in RF output. Q4 is a standard PNP keying switch, as discussed earlier. It not only actuates oscillator Q1, it also triggers dc switch Q5, which in turn activates Q6. When the key is closed, the signal energy to the receiver is shorted to ground by Q5. At the same moment, the series diodes, D2 and D3, are turned off by virtue of transistor switch Q6 being in the OFF state. This prevents signal energy from passing through the diodes to the receiver. When the key is up, Q6 conducts and provides a dc ground return for the diodes, which

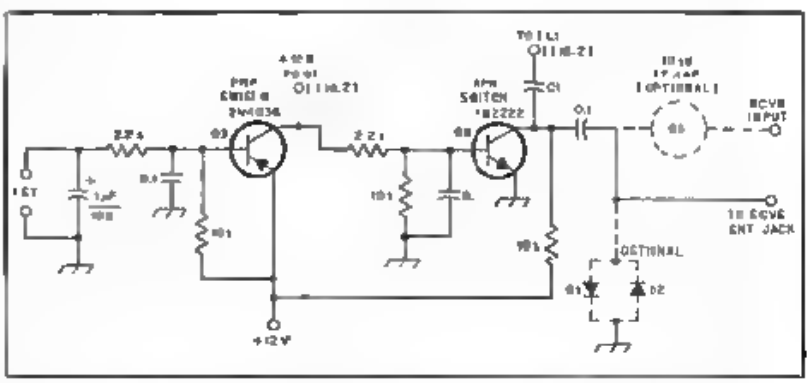


Fig. 8 — An improved TR system in which Q4 shorts the RF energy to ground when the key is closed. D1 and D2 are optional. They may be added as safety backup for Q4. Signal loss on receive is common with this simple TR circuit (see text), so an RF amplifier can be added at Q5 to compensate for loss in the TR circuit.

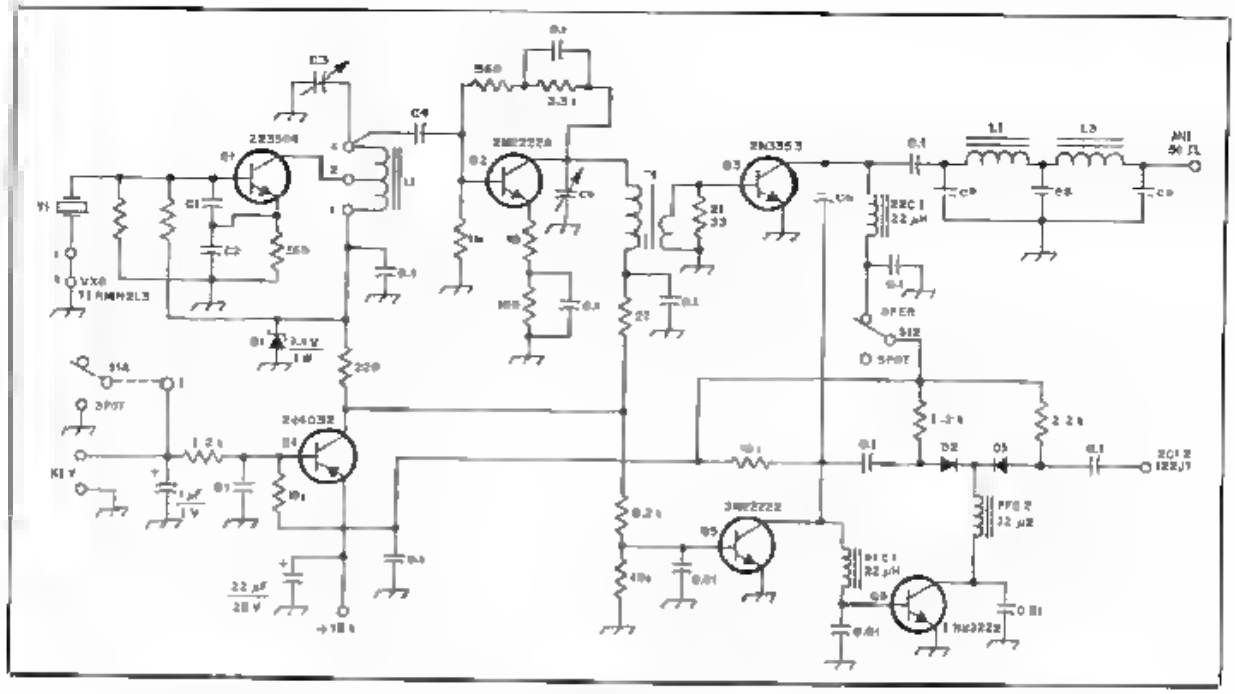


Fig. 9 — A somewhat elegant TR switching method that uses a series diode switch and a signal shunting transistor (Q5, D2 and D3). An explanation is given in the text. This is purely an illustrative RF circuit from Q1 through Q3. Therefore, there is no parts list.

enables them to reach the ON state. In effect, the action of Q5, D2 and D3 offer added attenuation and double protection for the receiver during transmit periods. The use of Q5 alone greatly reduces the signal level to the receiver input, as compared to the circuit of Fig. 7. I have measured only a few millivolts peak to peak from the receive line to ground when using a transistor shorting switch in place of TR diodes. There is no parts list for the transmitter of Fig. 9, since it is purely an example of how the switching techniques might be used. If you like QSK operation, you may want to experiment with this circuit.

In Conclusion

The intent of this primer on electronic switching is to stimulate thinking on your part, and to encourage you to work with semiconductor switching circuits. Since ICs contain diodes and transistors, many of them are applicable to circuit switching. A number of logic ICs are designed expressly for switching use and for gating.

Certainly, solid-state switches lend themselves well to use in compact circuits. The overall cost of a switching circuit may be somewhat less than that of a similar circuit containing mechanical switches. Also, by replacing relays with semiconductor switches, we can greatly reduce the current required by the overall circuit in our gear.

PIN diodes are designed especially for switching in RF voltage lines. An excellent example of PIN diode TR usage is given in *The Radio Amateur's Handbook*. See the chapter on keying (in recent editions).

You should have no problems in obtaining suitable switching devices these days. Ham radio flea markets and surplus outlets offer a plethora of diodes and transistors for this job, and the unit prices are often less than 10 cents! Perhaps it's time for you to "switch" to solid state!

From October 1987 QST, p 36:

Reducing AM Detection in Direct-Conversion Receivers

☐ While building equipment for the 40- and 30-meter bands, I discovered that AM detection is a common problem in D-C receivers. I used a singly balanced, four-diode detector followed by 85 dB of audio gain and a conventional RC active filter with additional gain. When the receivers were completed, both would detect any AM signals above about 200 μV in level. This is a problem because there are many such signals in the neighborhood of our 30- and 40-meter bands.

I went to some lengths to decouple and shield each receiver's LO, and to provide RF decoupling between the detector and the audio amplifier. Neither of these changes made any improvement.

Oscilloscope display of the detected AM signal showed an interesting peculiarity: At the receiver input, most signals exhibited symmetrical noise—but the *detected* AM signals showed only *negative-going* noise. This led me to suspect that the detection was actually taking place in the audio amplifier. Further, working with a receiver with no front-end selectivity, I found that sensitivity to AM detection decreased with increasing separation between LO and AM signal frequencies. This strengthened my hunch.

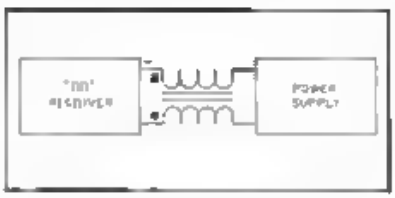
I solved the problem by installing a passive L-network filter, with a bandwidth of several hundred hertz, between the detector and the audio amplifier. I used a design similar to that shown in Fig 12 on p 77 of *Solid State Design for the Radio Amateur* with good results. With the filter installed, the modulation on AM signals of several thousand μV is inaudible with a 10-kHz LO/signal spacing.—*Dexter Bramwell, K7OWJ, 3139 Royalton Heights Rd, St Joseph, MI 49085*

From July 1977 QST, p 51:

Common-Mode Hum in Direct-Conversion Receivers

A direct-conversion receiver may be virtually impossible to use with an line operated power supply, owing to excessive hum. Part of this problem is that a direct-conversion receiver obtains most, if not all, of its gain at audio frequencies. Hence, the high audio gain makes the system subject to the smallest ac hum on the power supply. The cure for this problem is just better regulation in the power supply, which is easily realized with an integrated-circuit regulator.

A more subtle form of hum is also common and does not depend upon power-supply regulation. This hum is not present when no antenna is connected to the "dc" receiver. However, when an antenna is attached, a very rough sounding hum-like noise is heard. The amplitude of this response peaks in the antenna trimmer is heard. There are a number of possible explanations. The most realistic is that local-oscillator energy from the dc receiver is coupled into the power-supply lines. This energy is transformed back through the power supply when it is modulated by the rectifier diodes. The resulting hum-modulated noise is now coupled into the ac line. This signal is radiated and picked up by the nearby station antenna. Only the sidebands are detected.



Using a toroid with bifilar winding to reduce hum

While diagnosis of this problem may be subtle, a solution is deceptively simple and is shown in Fig 1. A large ferrite toroid is wound with a bifilar winding of reasonably large wire diameter. Ten turns of no 18 are usually suitable. The core is not critical although it should have a high permeability. An Amidon FT-82-75 is recommended. The effect of this balun-like circuit is to present a high impedance for any rf paths between the receiver and the power supply. Only the dc difference voltage from the power supply is applied to the receiver.

In the writer's station this method was applied with three different direct-conversion receivers. In two of the three cases the receivers were previously noise-plagued with battery power sources. With the toroid no differences could be detected when switching from a battery pack to a well-regulated ac supply.—*Wes Hayward, W7ZOI, 7700 SW Danielle Ave, Beaverton, OR 97005*

From September 1989 QST, p 38

Series-Resonant Circuit Enhances Desired Signal in QRP Rig

☐ During cut-and-try construction of a QRP CW rig that uses push-push doubling to produce 14-MHz drive from a 7-MHz VFO, I discovered that the stages following the doubler had output everywhere *except* 14 MHz! I solved this problem by installing a series-resonant tuned circuit between the doubler and its buffer stage (Fig 3). I have also successfully used series-resonant circuits between the antenna and output stage of monoband rigs to minimize TVI. (By the way, I first submitted something for Hints and Kinks in *1932*, but *QST* didn't publish that hint. I have since recovered from my feeling of injection and decided to try again!)—*Bob Kuehn, WØHKF, 1871 Silver Bell Rd, Apt 313, Eagan, MN 55122*

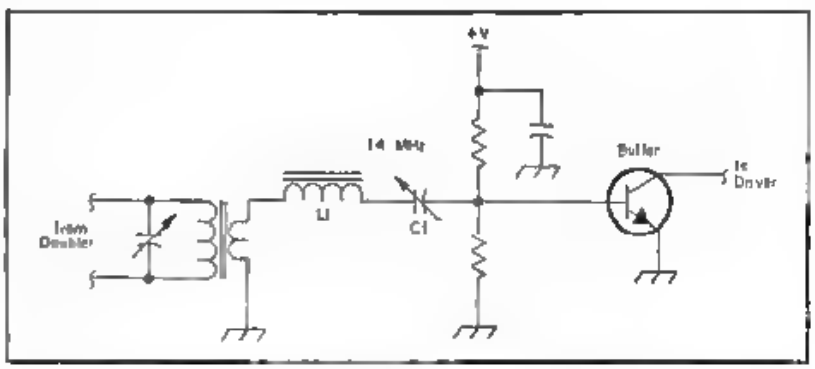


Fig 3—Bob Kuehn added this 14-MHz series-resonant circuit (L1C1) to clean up the output of a push-push doubler in his homemade QRP transmitter. L1 consists of 44 turns of no 24 enameled wire on a T-68-2 powdered-iron toroidal core. C1 is a small air-dielectric capacitor capable of being set to about 11.5 pF.

When radio amateurs first began using tube transmitters, the race to work the most miles per watt was on. In the '50s, transistors added a new dimension to QRP (low power) operating. And with today's ICs, it's possible to put together a complete station that fits into the corner of a knapsack: backpack into the wilderness, and enjoy worldwide communication!

QRP operating is fun. The equipment is generally simple and easy to build, but often performs like more sophisticated commercial equipment. Imagine the sense of accomplishment you'll get from operating equipment you built yourself. Some QRP Field Day stations operate a full 27 hours on a car battery—it's the perfect equipment for emergency communication when the power fails.

This book is a collection of projects published in ARRL publications over the past 15 years. Find out how to build receivers, transmitters, transceivers and accessories. There's a chapter on portable antennas. Power supplies, and a host of accessories are described. The chapter on design hints covers amplifiers, matching networks, electronic switching and direct-conversion receivers.

Are you looking to add an exciting aspect to your Amateur Radio interests? Come join the fun—give QRP a try!

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